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ENCYCLOPEDIA
ON
CATHODE-RAY OSCILLOSCOPES
AND
THEIR USES

by

JOHN F. RIDER

and

SEYMOUR D. USLAN

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FOREWORD

To the best of our knowledge this encyclopedia is the first attempt to present in one volume a cross section of cathode-ray-oscilloscope theory and applications which embraces virtually all the field of activity where this device is used. It is a far cry from our original text, "The Cathode-Ray Tube at Work," which appeared in 1935 and since that time has been reprinted thirty-three times. The present book is far more expansive and almost six times the size in words and even more than that in coverage.

We can recall very vividly the appearance of the first breadboard version of the cathode-ray oscilloscope in the early 1920's and the first commercial devices which appeared in the early 1930's. For a long time thereafter it was considered the exclusive tool of the research engineer and was very much limited to the laboratories of the electrical and radio equipment manufacturers. Even radio engineers found it a little confusing at the beginning, so it is not surprising that when it was offered to the maintenance branch of the radio industry it met with only half-hearted interest. Some of the servicemen of the nation purchased equipment but it did not enjoy the popularity which it rightfully deserved.

Today the capabilities of the device are fully recognized. The awe in which it was held for a long time has given way to an appreciation of its usefulness and the realization that it not only can do a better job in myriad fields than virtually all other test equipment, but, in addition, can lead to accurate, informative conclusions. Owners and prospective buyers have more respect than ever for the device and also feel that its inner workings may be understood, that it is not the incomprehensible device which it seemed to be years ago.

With the advent of World War II, especially the appearance of radar, the cathode-ray tube came into its own. As a display element in these equipments it became familiar to many individuals who previously never came in contact with electronic devices. At the same time, it also became better known to those military personnel who in their civilian capacities were actively engaged in the electronic field. In addition many of these individuals came in contact with cathode-ray oscilloscopes, since they were very much in use in connection with the maintenance of radar equipments.

Then with the cessation of active shooting in World War II and the rebirth of commercial television (it existed during 1939-1941), the commercial use of the cathode-ray oscilloscope was increased, for as television receivers spread across the nation, so did the cathode-ray oscilloscope. It was a vital link in the chain of maintenance operations. Of the many kinds of test equipment which can find utility in television servicing, the cathode-ray oscilloscope is without question one of the two most important ones, if not the more significant of the two. The other device we have in mind is the vacuum-tube voltmeter.

Concurrently with the increased use of cathode-ray oscilloscopes as a part of electronic equipment maintenance, there also took place rapid advances in the application of the device to electronic equipment employed in nonelectronic fields. The aircraft manufacturer, the industrial equipment manufacturer, the medical equipment manufacturer, and the educator all saw uses for the cathode-ray oscilloscope as a device which enabled more detailed analysis of actions. These stemmed from the development of devices known as "transducers" whereby nonelectrical phenomena are translatable into electric voltage and thus become suitable for application to the cathode-ray oscilloscope for visual analysis. Such actions as vibration, pressure, rotary motion, heat, light, and others now can be converted into voltage and the equivalent signals displayed on the screen of the cathode-ray tube used in the oscilloscope. The information gathered in this way has been of inestimable value in all forms of research — electrical, medical, industrial, atomic, — civilian and military.

Today cathode-ray oscilloscopes are the most versatile of all measuring devices because they enable display of waveforms over a wide frequency range. In the early days of the equipment it was felt that a top frequency of 100 kc was high. Today regular commercial apparatus are designed for performance at 100 Mc and the design details of oscilloscopes for operation at 10,000 Mc have been disclosed. With the rapid strides being made up the ladder of frequencies in the electromagnetic spectrum, the capabilities of response of the electron beam will be utilized more and more.

In view of its tremendous breadth of usefulness, it is understandable that a single volume such as this, devoted to the cathode-ray oscilloscope, cannot fully embrace every possible use or every possible variation of the device. We have attempted to cover the most significant applications in terms of the greatest amount of use. World-wide technical literature has been explored to gather the application content of this book. About 400 different technical publications from the world over arrive each month at the office of the publishers. These were reviewed carefully for information vital to the contents of this encyclopedia. While we know that many items have been omitted, those which were considered pertinent to the possible readers of this book were included. Maybe not in full detail, but sufficiently to enable recognition and correlation.

Equal weight has not been accorded every subject covered in this volume. It could not be done inasmuch as the purposes of the book demand more extensive coverage of some subjects than of others. It is possible that some readers who may have a special interest in some subject may find that it has limited coverage. Let us hope that the bibliography contained in this book will furnish them with additional sources of related information.

We realize, for example, that the study of power supplies deserves more space than was allotted to it, although what has been given conforms with the general purpose of this text. The same may be said about the testing of transmitters; the analysis of amplitude modulation is relatively complete for a variety of transmitters although all are of the high-level modulation variety, except the f-m systems. The latter are the subject of study at the present time and while the oscillograms presented in the book are not too complete, we feel that they are the first to be shown in any text of this kind. More will appear later when the book is revised. These oscillograms will serve, at least, as a starting point in this kind of work. It is hoped that the details concerning transmitter testing which were selected from other publications will satisfy the needs on special systems which are not fully covered by our own experimental investigations.

In connection with the above it might be well to remark that the fundamental facts contained herein have been written in such a manner as to make the cathode-ray oscilloscope understandable to all those who have occasion to use it at present, those who may have occasion to take advantage of its numerous salient features, and to those who may wonder about its potentialities in activities which have, heretofore, not made use of this extremely versatile electronic device.

As one glances through the contents of this encyclopedia, the attempt to develop the basis of operation will be seen readily. The only assumption made in this book is that the reader has a general background in electricity and magnetism and is in possession of the rudiments of vacuum-tube theory and operation. Such men are students in schools where radio and television are taught, the different schools which serve the communication needs of the Armed Forces and those individuals who have not had extensive electronic background, but are, nevertheless, interested in the theories underlying the operation of the oscilloscopes they are working with.

Understandably, the greatest emphasis has been given to electronic uses throughout the book and this material occupies about one-third of the book. The first two-thirds are devoted to those details which will enable any individual who has a cathode-ray oscilloscope to understand the device thoroughly so that if the success of his duties determines his dollar income, he is enabled to operate more rapidly and effectively with the minimum expenditure of time and effort, thus increasing his income. If, on the other hand, the reader of this book is contemplating the acquisition of a cathode-ray oscilloscope, the contents of this encyclopedia not only will make him familiar with all the details of the device, but, in addition, will enable him to select the one which most completely suits his financial capabilities and also his technical needs.

Then there is the student of electronics in whatever kind of school he may attend. This book is intended to be of material help to him by not only giving him the principles underlying operation but also fitting him for whatever device he may encounter when he completes his schooling, whether it be a civilian or a military pursuit. In this connection, it is hoped that the contents of Chapter 10 will be of material aid, for, while described as the "Basic Oscilloscope," it is a cross section of all kinds of oscilloscope devices, from the simplest to the most complex which is generally encountered in connection with electronic equipment operations, whether it be radio, television, or public-address equipment maintenance, radio transmitters, radar receivers, or those other equipments which may be modifications of these everyday devices as required to suit different purposes.

An innovation is attempted in one chapter of this book. This is the complex waveforms in Chapter 20. It is intended to serve those men and women who may have occasion to analyze equipment performance utilizing sine-wave voltage input to a system as a test signal and noting the output voltage waveform. Normally, such analysis requires the application of supplementary apparatus or the use of graphical analysis involving mathematics. In either case, it is a matter of time. An attempt is made here to save some of this time by showing a series of complex waves, all of which have been developed on a complex-wave synthesizer designed and built in the John F. Rider Laboratories. Thus the composition of each complex wave is given in terms of the relative phase of the components and the amplitude of the components. In all, about 1,600 such waves are shown and they include various combinations of a fundamental and harmonics up to the seventh. Obviously, such presentation can be carried on almost to infinity in the number of components and the phase-difference steps. We have selected those harmonic combinations which are most frequently encountered and which are most significant. It is hoped that these oscillograms will be of benefit to research workers and to students of electrical engineering. If this proves to be the case and it is felt by those who read these pages that such complex-wave analysis should be expanded so as to cover the needs of the electrical and electronic industries, either the volumes of the encyclopedia which are to come will contain an expanded group of complex waves, or it may appear as a separate volume of perhaps 25,000 to 50,000 such illustrations, if not more. It all depends on the responses we shall receive from the readers of this book.

A complete chapter has been devoted to the schematic diagrams of a wide variety of commercial oscilloscopes and related equipment. About seventy-odd oscilloscope devices are illustrated schematically. Some of them are accompanied by circuit descriptions and these are intended to correlate the device with some of the modifications of the basic oscilloscope as described in Chapter 10. Naturally they also explain the circuits of the specific instrument. Relative to these circuits, only the unusual or uncommon ones are treated in this manner. For that matter, if the same kind of circuit appears in several instruments, only the first one receives explanatory comment.

Several more comments concerning the contents of this text are in order. The authorship of the different chapters is indicated in the Table of Contents. However, in some instances, the contents of the chapters are abridgements of material taken from different scientific journals and proceedings. These are shown as credit lines under the illustrations and as footnotes relating to the different paragraph headings. Our sincerest thanks are tendered herewith to the different organizations who so willingly co-operated with us and extended permission to lift text and illustrations from their publications.

It is conceivable that in a work of this kind, there will be some credit omissions. We tried to minimize these as much as possible by checking and rechecking. Yet we know that, in some few instances, credit lines were omitted. For this we are very sorry and humbly apologize. We hope to find each of these eventually and to correct them at some future date. The reason for stressing this point is that we are desirous of giving credit wherever it is due, especially in the light of the fact that all illustrations, photographs, and oscillograms which do not bear credit lines are our own, having been made by our own art department or photographed in the John F. Rider Laboratories. We reiterate that we regret the inclusion of drawings and photographs among our own which rightfully should be credited to others. Concerning the black-on-white line drawings of the complex waveforms in Chapter 20, each of these is a reverse print of an actual oscillogram photographed directly from the screen of a cathode-ray oscilloscope. Few if any oscillograms were drawn; by far the vast majority of such illustrations throughout the book were actually photographed and are shown as they appeared on the screen of the cathode-ray tube.

It may be well to comment on the fact that the presence of credit lines under different photographs and illustrations of commercial oscilloscopes is not intended to create any impression that these are the preferences of the authors relative to commercial equipment. All manufacturers were solicited in like manner for information concerning their products and whatever else they had to offer. Some responded much more readily than others and an omission is simply evidence of failure to receive responses to our communications and should not be construed in any adverse manner as reflection either on the manufacturers or on their products. Then again some data arrived after the book was at the printer and could not be included because the forms which contained similar material already had been printed.

While we are anxious to give all organizations equal recognition and thanks, we must of necessity accord our heartfelt gratitude to the Allen B. Du Mont Laboratories, Inc., to the RCA-Victor Manufacturing Co., Inc., and to the Sylvania Electric Products Co., Inc., for their exceptional willingness to go out of their way to prepare special data for inclusion in this book.

We wish to take this opportunity to express our thanks to the members of the Art Department, the typists, and the editors of John F. Rider Publisher, Inc. for their co-operation in accomplishing this book, especially Henry Jacobowitz, and Norman Weinstein for checking technical accuracy. It turned out to be very much bigger than was originally contemplated and this meant that everyone was pushed to the limit in the attempt to meet a publication date. We had hoped to complete the work in much less time than actually was consumed, but the rapid growth among the manufacturers of cathode-ray oscilloscopes, the many different kinds of oscilloscopes which have become commercial products, and our desire to include the greatest number were reasons for the delay. We hope that those who have been waiting patiently for the appearance of this encyclopedia will feel that it was worth while waiting for.

September 1950
New York

JOHN F. RIDER
SEYMOUR D. USLAN

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CHAPTER 1

INTRODUCTION

This is the story of the cathode-ray oscilloscope and its applications. What is a cathode-ray oscilloscope? In answer we might say, "It is a device which can portray electrical phenomena visually." This may seem to be a very limited definition, but it is actually extremely broad in its implications.

Oscillograph, Oscilloscope, and Scope

Sometimes, various names, such as "low-frequency oscillograph," "video wide-band oscillograph," "wide-band, high-gain oscilloscope," "projection oscilloscope," or simply "oscillograph" or "oscilloscope," are used. Some of these names indicate special performance characteristics, while others are merely identifications of the uses of this extremely versatile device. The three most popular are oscillograph, oscilloscope, and the abridged version of the second name, the scope. All mean one and the same thing, although the name "synchroscope" has a more precise meaning. It is an oscilloscope, but intended specifically for the measurement of pulses and transients. It is, however, included in the general category of the oscillograph.

In this book, the word oscillograph will be replaced by oscilloscope because it is a more popular name, and more appropriate to the device. It has the same meaning as its abridgement "scope," which we do not use only because it seems more fitting to be formal in the text.

Electrical Phenomena

Practically speaking, electrical phenomena are voltage and current. Because the two are proportional, whatever is said about voltage applies equally well to current. Such details of a voltage as instantaneous amplitude, phase, frequency, waveform, composition, and duration, all are determinable, within-certain frequency limits, by means of a cathode-ray oscilloscope. The origin of the voltage is immaterial; it may occur in any type of circuit or in connection with any component, L, R, or C. The circuit involved may be a part of a complete system or it may be a complete system in itself. The suitability of a specific oscilloscope to a particular application may be determined by comparing its performance specifications with the requirements of the problem.

The cathode-ray oscilloscope may be used with equal facility for checking transmitters, receivers, amplifiers, and any apparatus embracing these elements. It can be used for measurement of operating potentials, for the adjustment of tuned circuits, or for the determination of the operating conditions within these systems. Within its range of utility are the comparison of the behavior of one electrical device with that of another, the calibration of such devices, and the study of their electrical characteristics under varying conditions. A voltage may represent an amplitude-modulated, frequency-modulated, television, or any other kind of signal. It may be noise, artificial or natural—the origin of the signal is unimportant.

The uses of the oscilloscope are not restricted to electronics as applied to communications; rather it will serve any electronic system. It is suitable for application to power equipment, to light and heavy electrical machinery, and to all industrial electronics systems as well.

Assuming the existence of a suitable *transducer,* that is, a device which will convert a mechanical or any other nonelectrical phenomenon into an equivalent electrical voltage, the cathode-ray oscilloscope will serve admirably for the investigation of all physical quantities. Distance, light, temperature, pressure, sound, magnetic effects—all these and many others are phenomena which can be observed, studied, and measured by means of cathode-ray-oscilloscope equipment. Thus the device has universal application in science, industry, medicine, and education. It can serve all these masters and serve them well.

So much for definition and applicability. The stage has been set and we can procede with the rest of the story.

Early Types of Cathode-Ray Tubes

During the years 1921-1924 several publications[1,2,3] announced a new type of tube. It was a hot-cathode, low-voltage cathode-ray tube, and it was a part of a

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[1] Johnson, J. B., "A low voltage cathode ray oscillograph," *Phys. Rev.*, vol. 17, pp. 420-421, 1921.

[2] Johnson, J. B., "A low voltage cathode ray oscillograph," *Elec. Commun.*, vol. 1, pp. 57-62, November, 1922.

[3] Kipping, N. V., "Wave form examination with the cathode ray oscillograph," *Elec. Commun.*, vol. 3, pp. 69-75, July, 1924.

system which could conveniently visually portray sound patterns and many varieties of electrical and magnetic phenomena. As might be expected, it originated with that engineering organization which was so greatly interested in the transmission of sound, the Bell Telephone Laboratories, the development branch of the American Telephone and Telegraph Company.

By that time radio communication was a reality. Commercial broadcasting was several years old, the radio amateurs had explored the possibilities of the short-wave region of the electromagnetic spectrum, and the vacuum tube had been in use for almost two decades. The range of frequencies used for communication purposes had expanded greatly. The numerous problems associated with the equipment utilized also had multiplied. Research was accelerating in all directions, and any new tools which might aid in this activity were welcomed by the industry as a whole.

Instruments for the examination of electrical and magnetic phenomena had been known for quite some time. The electromechanical oscilloscope, better known as the "string" oscilloscope because of its construction, had been conceived as early as 1891. With a modification by Duddell in 1893, this oscilloscope set the pattern for most of these devices in the future.[4] The basic unit consisted of a pair of parallel wires suspended in a strong uniform magnetic field. A tiny mirror was cemented to these wires at their midpoint in the field. The current to be observed was sent through these wires, traveling in one direction through one, and in the opposite direction through the other. The fields due to the current interacted with the steady field and caused a vibratory motion of the mirror corresponding to the instantaneous changes in the magnitude and direction of the current. A light source projected a beam of light onto the vibrating mirror and the reflected beam, in turn, was projected onto a moving strip of photographic film. The direction of motion of the film was parallel to the axis of vibration of the mirror, and by moving the film at a constant rate, the time base was formed. After development, the film was a permanent record of the instantaneous variations of the current with time, similar to modern photographs of current and voltage waveforms.

Obviously such a system had numerous limitations, foremost of which was the restricted frequency range due to the inertia of the vibrating system. The natural frequency of the vibrating system was made as high as possible, but even then the useful upper-frequency limit was very low in comparison with the range of frequencies which required investigation. Another deficiency was the user's inability to view the action instantaneously. Passing years witnessed numerous improvements. The frequency range was increased, although it never exceeded the upper limit of the audio range. Instantaneous visual observation was accomplished by means of a system wherein the vibrating mechanism was linked with a pen recorder which traced the pattern of vibration onto a strip of paper which passed the pen at a constant rate. Such devices are in use today, and, understandably, the frequency range is even more restricted because of the increased mass of the recording assembly.

Another means for accomplishing visual observation of the action in the early days of the electromechanical oscilloscope was an elaborate optical system. The vibrating oscilloscope mirror reflected the light beam onto a second mirror, which either rotated at a constant speed or vibrated at a constant rate, its axis at right angles to that of the oscilloscope mirror. The second mirror was the time-base mirror and the light beam reflected onto a screen traced a pattern which represented both amplitude and time—since the beam of light was displaced vertically and horizontally simultaneously.

Such mechanical oscilloscopes frequently contained as many as four, five, or even six elements so that several electrical quantities could be observed at the same time. These elements could be connected in series with circuits for current observations and across circuit elements for voltage observations. Strangely enough, fairly high orders of sensitivity were available, and as cumbersome as this description makes the device appear, numerous portable units were devised as time passed, and modern versions are available today.

As we stated earlier, the range of frequencies to which these units were applicable was limited. Phenomena of high frequency or very short duration were beyond their capabilities entirely. When the cathode-ray tube was introduced by Johnson, he assigned to it the name "Braun tube" in honor of the individual who invented the basic principle of its operation in 1897. It was presented as a part of a cathode-ray-oscilloscope device which was capable of response at frequencies up to several hundred thousand cycles, and it aroused tremendous interest in the engineering fraternity.

The world was not without a cathode-ray oscilloscope prior to 1921. The word "oscillograph" had been in the English language for a comparatively long time; it was defined as "an apparatus for recording or indicating alternating-current waveforms or other electrical oscillations."[5] When the term was conceived, it applied to the "string" oscilloscope because this was the only such device extant.

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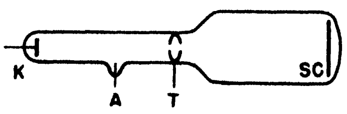
[4] Duddell, W. D. B., "Oscillographs," *Electrician*, vol. 39, p. 636, 1897.

The cathode-ray-tube device was of an entirely different character. It was a high-voltage tube operated at potentials as high as 50,000 volts, voltage sufficient to forcibly pull electrons out of a "cold cathode." These electrons were accelerated toward the other end of the tube, and in order to get there, they had to pass through a tiny aperture at the center of a shield or baffle. Those electrons which penetrated this tiny opening formed a relatively narrow beam, and struck a specially prepared screen which was located inside of the tube at the far end. The point of impact of the electron beam on the screen was indicated by fluorescence of the screen during the period of excitation. An elementary idea of the construction of this tube can be had from Fig. 1-1. The electron beam was found to respond to electric and magnetic fields and its behavior, while under such influences, could be noted by observation of the fluorescent screen. In comparison with the modern tube, the original Braun tube was very primitive, but no one can deny that a close parallel in fundamental behavior between this and the modern tube exists, as we shall soon see.

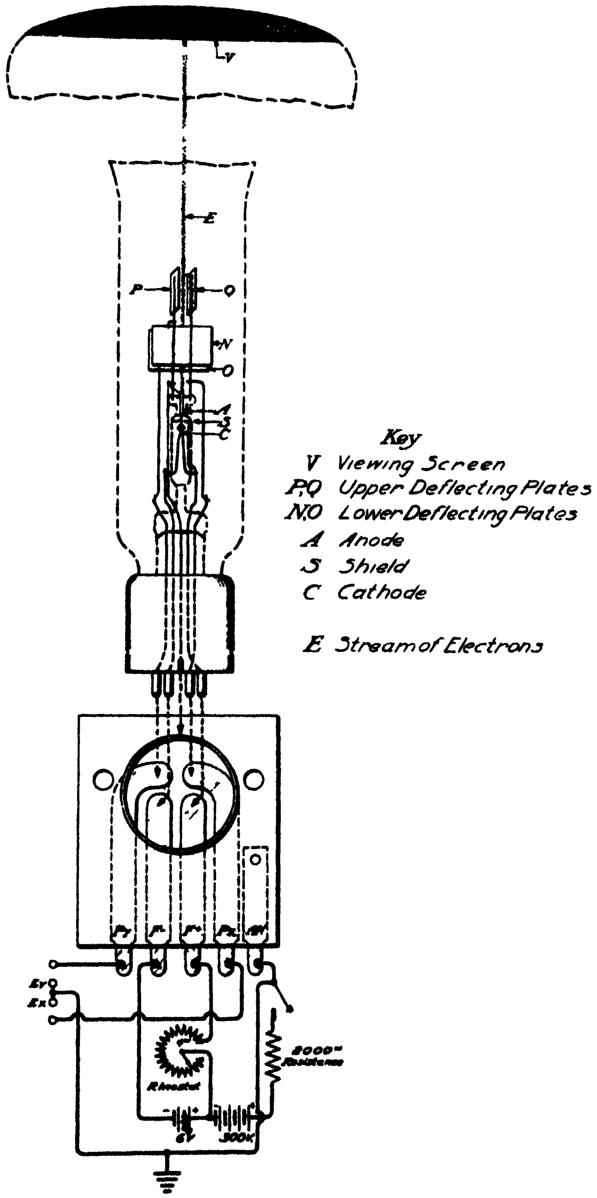
Fig. 1-1.—Schematic representation of the Braun tube.



The primitive system was subject to numerous modifications through which greater control of the electron beam was achieved, and, in time, the device enjoyed use in many industries, especially the power field. It was modified in such a way as to enable recording by placing a photographic film within the chamber so that it became the target for the electron beam. Under these circumstances, removal of air was necessary each time that a new photographic plate was inserted. As a device which utilized a cold cathode, very high accelerating voltages were involved in its operation. Analogous to it, although very much different in purpose and design, is the present electron microscope. Neither of these devices comes within the province of this text, and we mention them merely to illustrate the versatility of the cathode-ray tube.

The Hot-Cathode Cathode-Ray Tube

A chain of developments led to the Johnson tube (Western Electric 224A) which appeared in the early 1920's. When announced, it offered a convenient source of electrons which could be formed into a beam and accelerated toward the viewing screen at very low voltages. The virtually inertialess electron beam, which was not entirely composed of electrons in that model, was the vibrating mechanism, and it was capable of responding to relatively high frequencies. Because of the gas content of the tube, the frequency range was



Courtesy Western Electric

Fig. 1-2.—Pantographic view of Western Electric 224, showing external connections.

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[5] "Webster's New International Dictionary of the English Language," G. & C. Merriam Company, Publishers, Springfield, Mass., 1920.

limited to several hundred thousand cycles, and while this is very low by modern standards, it nevertheless was revolutionary for its time.

The tube was compact and simple to use, as were the systems devised for its application as an oscilloscope. It permitted instantaneous viewing of the action on a screen made of special materials painted onto the inner surface of the tube at the widest end of the pear-shaped glass envelope. This replaced the elaborate optical system necessary with the mechanical oscilloscope. If photography were necessary for permanent recording, it could be done by using an ordinary camera, by simply focusing onto the screen of the tube from the front. Naturally the film speeds and film sensitivities of today were not available then, so that when we remark that photography was simple, we mean relative to the processes involved in the other types of oscilloscopes. Today photography is no problem at all.

The electron beam in the 224A was placed under the influence of either one or two electrostatic fields by means of two pairs of deflection plates contained within the envelope. These were so positioned as to allow the electron beam to pass through the center of the space between the two plates of each pair, while advancing to the screen. The two sets of plates were oriented at right angles to each other so as to apply deflection forces at right angles to each other. Vertical deflection corresponded to the amplitude and the horizontal deflection corresponded to time, the latter requiring a voltage which deflected the spot horizontally at a constant rate.

The beam could be deflected electromagnetically by means of two pairs of coils, slipped over the narrow part of the envelope and located at the proper point along the electron trajectory. This placed the electron beam within the fields of the coils. A pantographic view of this tube is shown in Fig. 1-2. The electrostatic-deflection plates are visible in this illustration, and the relative location of the electromagnetic-deflection coils is shown in Fig. 1-3. Technically, the latter illustration has very little significance; it revives nostalgic memories of the cathode-ray oscilloscope the writer constructed in 1927.[6]

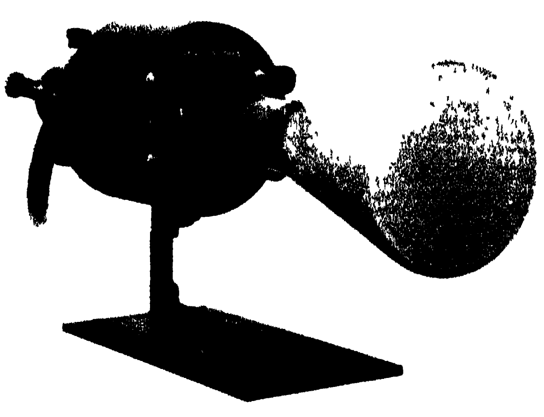

Fig. 1-3.—The Western Electric 224A cathode-ray tube shown with deflection coils.

A New Approach in Oscilloscopes

All in all, the hot-cathode, low-voltage cathode-ray-oscilloscope tube was a significant development. It was a new concept in oscilloscopes—the birth of a device which would be compact, and susceptible to myriad uses by the individual probing the depths of electrical and magnetic phenomena and circuits. The level of operating potentials which this device used opened new vistas of use, for it was a device which could be used where the actions under observation were at low levels. No longer was the cathode-ray oscilloscope restricted to certain specific industries where the phenomena to be observed involved very high potentials.

Admittedly, the low-voltage cathode-ray oscilloscope remained an awe-inspiring device for many years after its initial announcement. It saw only limited use in radio laboratories. It was not until the early 1930's that complete low-voltage cathode-ray oscilloscopes appeared commercially.[7] They differed from the original Johnson version in that the tubes used were of the high-vacuum type, which immediately widened the span of the frequency spectrum over which they could operate.

The original uses envisioned for the first hot-cathode cathode-ray-tube oscilloscope have been multiplied many times over, not only in the electronic industry, but also in many nonelectrical activities. In fact the electronic applications of even the modern cathode-ray oscilloscope are but a segment of the circle of its possible uses, as we shall soon see.

The Western Electric 224A is the direct antecedent of the modern tube. In appearance it greatly resembles the tubes we use today, as shown in Fig. 1-4. Its pear shape is duplicated in many different models of the present tube (A), thus showing that those responsible for the original conception of the tube had chosen wisely. The bottle shape (B) is another modern tube; its modified shape was necessitated by the incorporation of certain new features into the design of the tube.

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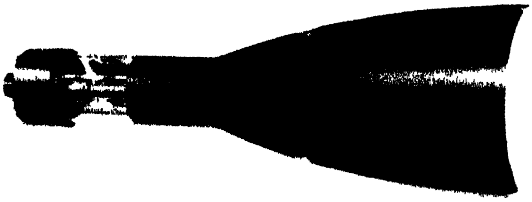
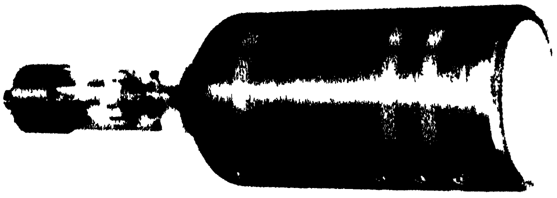
[6] Rider, John F., "The cathode ray oscillograph tube," *Radio Eng.*, vol. 7, p. 982, October, 1927.

[7] "A Catalog of Equipment for Oscillography," Allen B. Du Mont Laboratories, Inc., Clifton, N. J., 1949; p. 1.



(A)



(B)

Courtesy Du Mont Labs

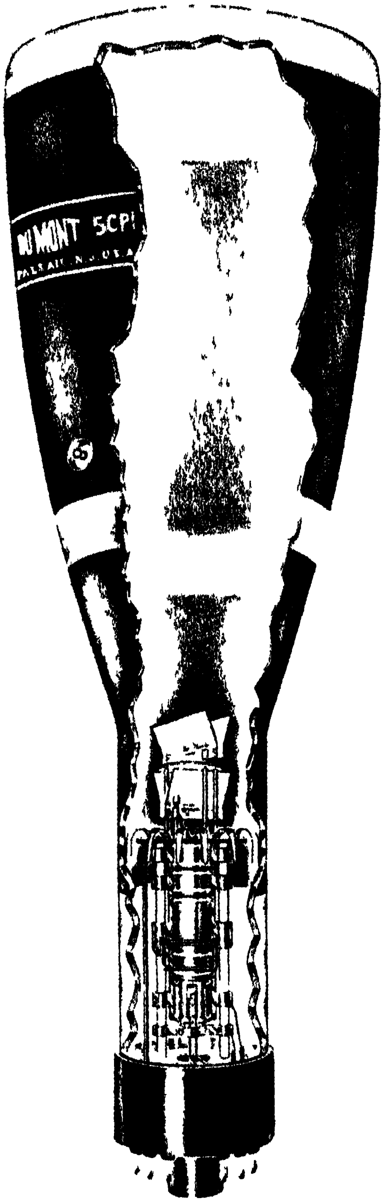
Fig 1-4—(A) Modern pear-shaped tube and (B) modified bottle-shaped tube.

We will see later that while, in general, the shape of modern tubes might be said to conform to these two distinct types, variations will be noted, but these arise from physical necessities introduced by the equipment utilizing them.

While we are comparing the old with the new, it is very interesting to note that the principle of operation has not changed very much. The physical design of the electrodes within the glass envelope has been modified, the consequences of research; but as to the fundamental internal organization, there is great similarity between the old and the modern versions, as shown in Fig 1-5. This is an example of the electrostatically deflected tube. The electromagnetic type is shown later.

A Research Tool

If we are to seek a reason why the cathode-ray tube has for so long been considered the tool of the research worker, we might go back to its recommended uses at the time of its debut. Its immediate utility in connection with the telephone industry and all electronic equipment was evident. The association was direct, and the cathode-ray tube soon became the key to the hitherto inaccessible recesses of electronic circuits. It showed what was happening and, in so doing, opened new paths of research to the scientist. The oscilloscope resulted in new systems and devices which, in turn, increased its own potentialities and gave man the knowledge necessary to invent many more new devices. It laid its own road to its own success.



Courtesy Du Mont Labs

Fig 1-5—Internal view of electrostatic Du Mont type 5CP1.

As a research tool, the value of the cathode-ray tube has increased manifold But it lost its identity as strictly the tool of the engineer. As a part of the oscilloscope, it has become a byword in electronic maintenance equipment of all sorts, in science, education, in all technology. It is a must for every television service shop.

The cathode-ray tube has become known to the public as the "picture tube" in a television receiver. Men from all walks of life who participated in the radar activities of the different branches of the Armed Forces are familiar with it. Now that peacetime applications of the "echo" principle of detection of unseen objects is fast becoming widespread, personnel on the high seas are acquainted with it. The student studying electronics uses it daily. Truly it is a wonderful device which serves science, industry, and numerous other fields of activity as a means of disclosing action which formerly could only be guessed at.

What Is the Cathode-Ray Tube?

What is this marvelous device which is capable of serving so many masters? On the surface, it may appear as a complicated device. In reality, it is not so—anyone with an elementary knowledge of electricity and magnetism can readily comprehend its action. In essence it is a vacuum tube—a special type of tube in which a very narrow beam of electrons behaves like an electronic "pencil" and "draws" a visible trace or pattern on a specially prepared screen.

Compared to the ordinary variety of receiving or transmitting type of vacuum tube, the cathode-ray tube is singular in many respects, but you will also note, as its theory of operation unfolds, that there are many parallels between its principle of operation and that of the general purpose vacuum tube. We make this statement in an effort to ward off any misconception that its understanding involves new theories and principles which are accessible only to the man with an engineering background. Such is not the case, even though the concepts of electron optics are relatively new ones to the individual whose work in electronics has been restricted to the ordinary vacuum tube.

Referring again to Fig. 1-5, the source of electrons is a hot cathode located at the base end of the tube inside of the envelope. An assembly of electrodes forms the narrow electron beam and accelerates the electrons toward the screen which is located furthest from the cathode, at the widest portion of the tube. The screen is a coating of luminescent material on the inside surface of the large, relatively flat end.

When the high-velocity electron beam strikes the screen material, it causes the atoms located at the point of the beam impact to luminesce. The result is a luminous dot about 0.005 to 0.05 inch in diameter, the diameter of the beam when it strikes the screen. Wherever the screen is struck by the beam, a tiny dot of light appears. If this little dot of light is made to move rapidly enough across the screen, a luminous line will appear. This leads to the statement that whatever trace or pattern or picture appears upon the face of a cathode-ray tube, it is made by a single dot which moves across the surface of the screen.

Deflecting the Electron Beam

To make proper use of the cathode-ray tube, the electron beam must be made to move across the surface of the screen. This is called deflection. As stated earlier, this displacement of the beam can be accomplished by either electrostatic or electromagnetic means, by placing the beam under the influence of the appropriate field. This is the purpose of the deflection plates in the electrostatically deflected tube and the deflection coils in the electromagnetically deflected tube. In both instances, the deflection system is located at appropriate points along the path of the beam. Figs. 1-6A and B illustrate deflection in the electrostatic tube.

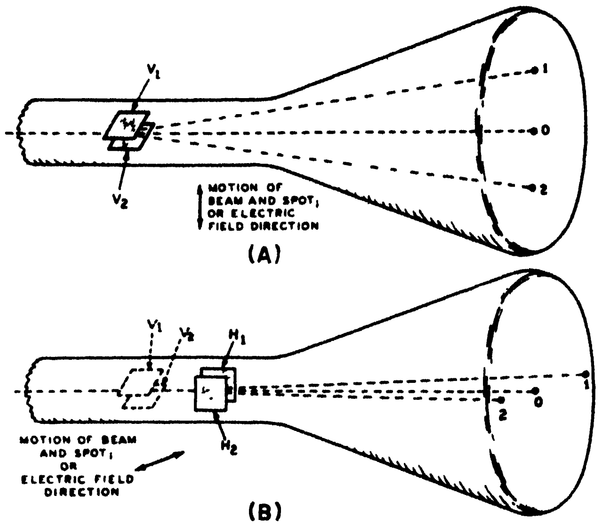

V1
V2
MOTION OF BEAM AND SPOT, OR ELECTRIC FIELD DIRECTION
1
0
2
(A)
V1
V2
H1
H2
MOTION OF BEAM AND SPOT, OR ELECTRIC FIELD DIRECTION
1
0
2
(B)

Fig. 1-6.—Deflection in the electrostatic tube (A) illustrates vertical deflection by plates V_1 and V_2 and (B) illustrates horizontal deflection by plates H_1 and H_2.

The dotted lines in Fig. 1-6A represent three paths of the electron beam. Without any deflection voltage applied across the vertical-deflection plates, the point *0* results. With deflection voltage of such polarity as to move the beam toward the upper limit of the tube screen, we get point *1*. Then, deflecting the beam toward the lower limit of the tube screen, we get point *2*.

Analogous conditions associated with the horizontal-deflection plates are indicated in Fig. 1-6B. These plates are vertically oriented, and the three points of impact of the beam are: *0* for no deflection, *1* for deflection toward the right, and *2* for deflection toward the left. The dotted rectangles within the tube envelope in Fig. 1-6B are the vertical-deflection plates.

Electrostatic deflection is the result of either the force of attraction between unlike charges, here the negative charges in the electron beam and the positively charged deflection plate, or the force of repulsion between like charges, between the negative charges in the electron beam and the negatively charged deflection plate. When the deflection voltage is applied to the deflection plates, one becomes positively charged, naturally, and the other assumes a corresponding negative charge.

Electromagnetic deflection is the result of the interaction of magnetic fields—the field which surrounds the moving electrons which comprise the beam and the field developed by the deflection current in the coils.

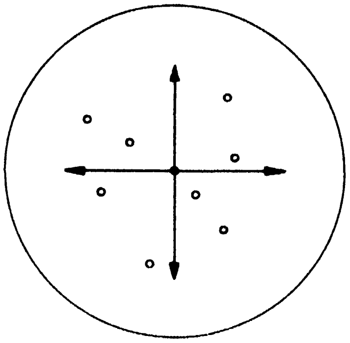


Fig. 1-7.—Beam positioning; the center dot being the undeflected beam, and the small circles indicate positions of the beam under various steady-state deflection conditions.

A representation of the two perpendicular directions of deflection is given in Fig. 1-7. The heavy dot at the center is the location of the undeflected beam. Under ideal conditions, it is located at the geometric center of the tube. Since the two deflection forces may be applied simultaneously, the electron beam may be made to strike the screen at any point on its surface, the exact location being determined by the relative intensities and directions of the individual deflection fields. A few positions of the beam under various steady-state deflection conditions are indicated by the small circles between the crossed arrows.

Under any one set of operating conditions, a linear relationship exists between the strength of the deflection force and the magnitude of deflection. Since the strength of the electrostatic field is proportional to the deflection voltage which causes it, and the strength of the electromagnetic field is proportional to the current which causes it, the movement of the dot on the screen is directly related to the instantaneous changes in the deflection voltage or current, whichever is responsible for the deflection field.

An explanation of the reason for the two deflection forces acting on the electron beam at right angles to each other might be in order here. All action takes place in time. In order to show properly the manner in which an electrical quantity varies, it must be shown in relation to time. A 60-cps voltage completes a cycle of amplitude variation once every 1/60 of a second. To portray a current or voltage waveform or any other quantity, varying or otherwise, the passage of time must be indicated on one axis of the figure.

By arranging a deflection which moves the beam at a constant rate in a horizontal direction, and by permitting the vertical deflection to correspond to the instantaneous variations in the amplitude of the waveform being investigated, the resultant motion of the beam will trace out a pattern which indicates instantaneous amplitude relative to time, as shown in Fig. 1-8.

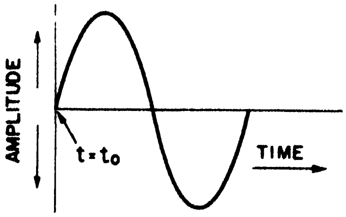


Fig. 1-8.—One cycle of amplitude variation along time axis.

The above explanation should not be construed to set rigid boundaries around the use of the horizontal-deflection system. Numerous applications are possible in which the horizontal deflection does not indicate time. The one we have given is fundamental to the display of waveforms in time, however, and should be accepted as such. Other applications which might use the horizontal deflection in a different way will appear in subsequent pages of this book.

The Path of the Electron Beam

The result of an abridged description of the deflection system might lead one to believe that changing the path of the electron beam by means of deflection fields will interfere with the beam's advance to the screen. This result is possible, but only when the deflection field intensities are excessive. Under normal conditions of operation, the deflection fields simply change the path of the beam and do not interfere with its advance to the screen surface.

Concerning the movement of the beam across the screen, it can be made to follow any desired path. It can sweep across the screen once in a prescribed period of time, it can do this at preset intervals, or it can be made to retrace its path as frequently, and for as long, as the operator may desire. It can do this slowly or very rapidly, although the appearance of the dot of light on the screen when the beam moves very slowly or rapidly calls for certain specific characteristics of the tube, especially the material used for the screen. We need not dwell upon this feature at this time, since it is treated in Chapter 7.

Frequency Response of the Electron Beam

By frequency response, in this case, we mean the ability of the beam to follow changes in the intensity of the deflection fields. All things being equal, the beam can respond to frequencies far beyond obtainable deflection frequencies. In order to utilize this response, it is necessary that special designs be employed inside the tube. Success has been achieved in this direction but, at the time of this writing, only to about 10,000 Mc. No doubt, in time, this range will be extended.

We may explain these frequency limitations briefly. The higher the frequency of the deflection field, the faster is the movement of the electron beam across the tube screen. This reduces the period of excitation of the atoms of the screen material, and consequently, the resultant light intensity on the screen. As the frequency is increased, the light intensity decreases, until the trace becomes useless, or invisible. To counteract this effect, greater accelerating potentials must be used. This increases the velocity of the beam, exciting the screen more strongly, and yielding greater light intensity. Greater accelerating potentials require, in turn, greater deflection voltages, as we shall see later, in order to deflect properly the faster-moving electrons.

Another factor is the finite velocity of the electrons which comprise the beam. These must pass through the deflection fields. If the polarity of these fields changes too rapidly compared with the time required for the electrons to traverse the field and be acted upon by it, the net deflection will be reduced or distorted, or no deflection may take place at all. This is a limitation discussed in Chapter 6.

Varied Density of the Beam and Spot Intensity

Before enumerating the many things the cathode-ray tube can do, it is imperative to mention another fundamental form of control of the beam. Since the electron beam is formed within the tube, means exist for the control of its density. Assuming any arbitrary velocity given to the beam electrons, the electron density may be controlled over a wide range so that, if desired, the dot of light visible on the face of the tube may be changed from normal brightness to invisibility. This is a very important control, for it is the means used to produce the picture seen on the television receiver screen. The deflection in the television picture tube is simply dot-positioning action, and the picture is the result of varying the light intensity from white to black at the different points on the screen. These light and dark points correspond to points in the scene being televised.

The above is not the only use for such intensity modulation of the beam; there are many others which are associated with other applications.

CLASSIFICATION OF CATHODE-RAY-TUBE USES

As a matter of general interest, it is possible to classify the uses of the cathode-ray tube into three major groups. These are:

1. Cathode-ray tubes used as display devices in receiving systems.
2. Cathode-ray tubes used as generators.
3. Cathode-ray tubes used in measuring equipment.

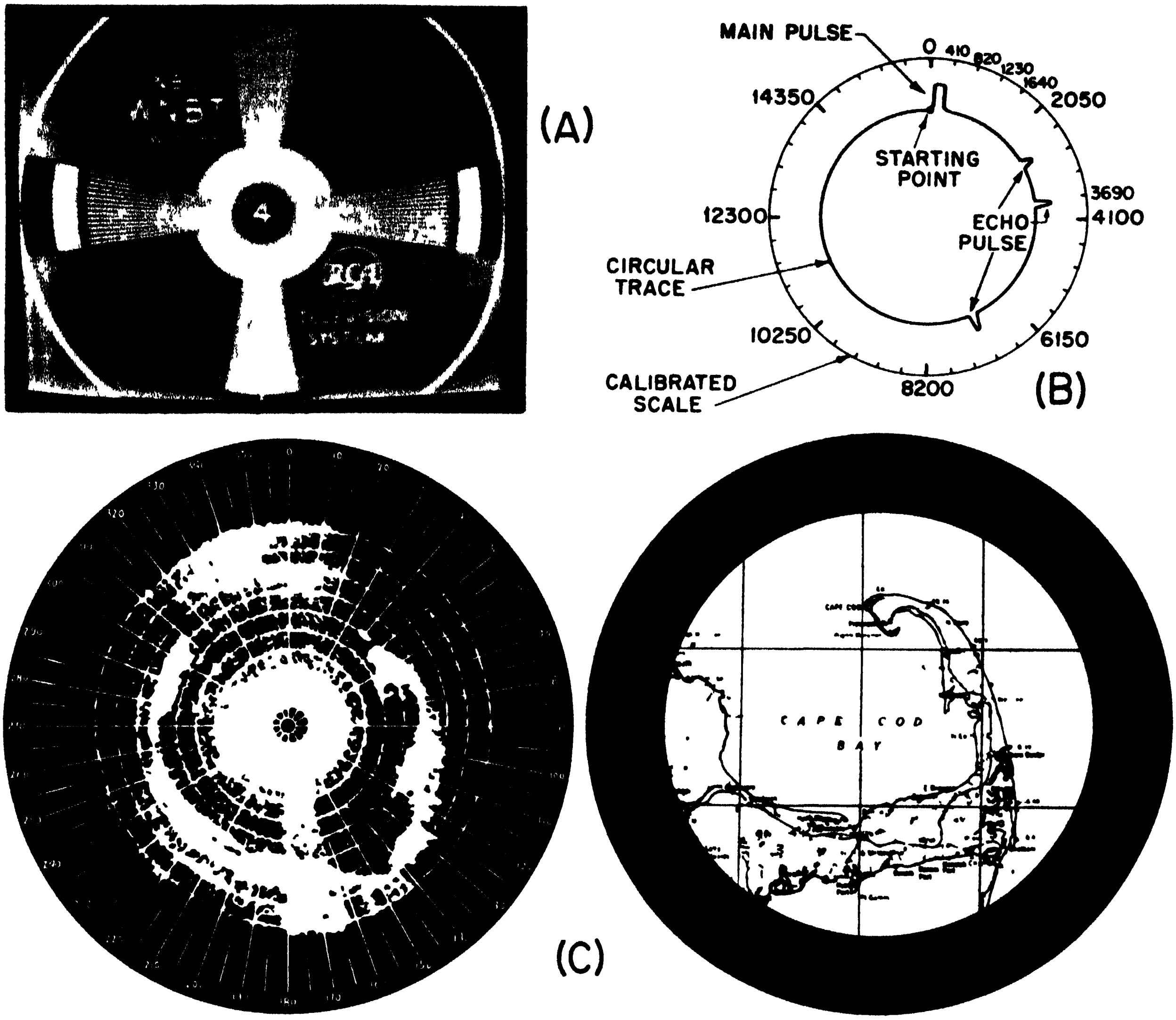
The Cathode-Ray Tube in Receivers

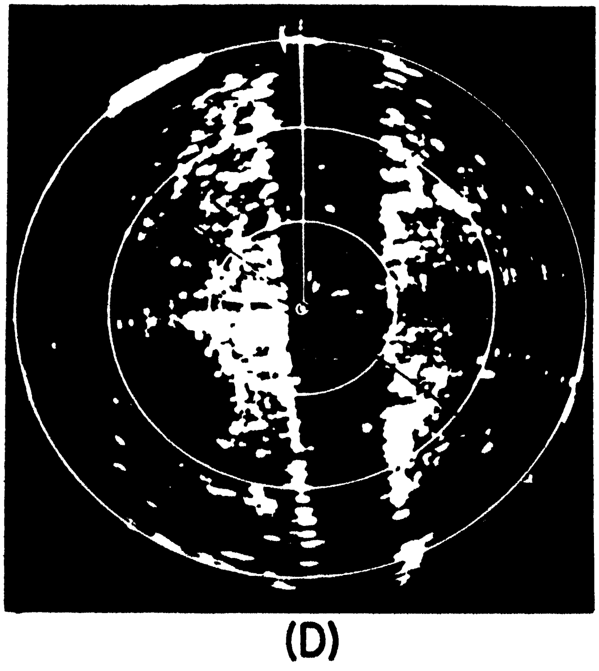
An appreciation of the first use is easy. Three and one-half million television receivers employ the cathode-ray tube either as the viewing screen on which the picture appears, or as the source of the picture which is enlarged for projection onto a separate viewing screen.

Another example is the radar receiver. A pulse is sent out into space from a transmitter. One or more target objects reflect some of this electromagnetic energy, and the reflections are picked up by a receiver adjacent to the transmitter. The reflected signal appears on the calibrated screen of a cathode-ray tube, and specific information concerning the location of the reflecting target, relative to the receiver, is determined from the trace on the cathode-ray-tube screen.

A type of radar equipment which is enjoying expanding peacetime application is the device which develops a map of the land or sea area being scanned by the equipment. In this system, a short-duration pulse is repeatedly transmitted from an antenna which is located above the area being examined. The antenna rotates azimuthally several times per minute, scanning the given area by small sectors. A reflected signal returns from each small sector, and, after the usual handling by the receiving equipment, appears as a spot, or a series of spots, of light on the screen of the cathode-ray tube. The net result is a map of the area scanned. A display device of this kind used for Armed Forces or civilian pursuits is called a PPI, or Plan Position Indicator.

Examples of the screen patterns described above are shown in Figs. 1-9 A, B, C, and D. The first (A) is a picture from a home television receiver. The second (B) is an example of what was seen by thousands of





(D) Courtesy Radiomarine Corp

Fig. 1-9—Typical screen patterns; (A) normal test pattern on a home television receiver, (B) radar detection of unseen enemy planes, (C) PPI radar map of Boston-Cape Cod peninsula, and (D) PPI radar map of the Hudson River.

ground-force radar operators during World War II who manned antiaircraft artillery radar equipment which showed the distance (range) of unseen enemy planes flying through the sky. The echoes from the targets appear as pips on the range sweep, at varying distances from the main transmitter pip.

The third (C) is a map of the Boston-Cape Cod peninsula as seen on the PPI tube screen of a radar-equipped plane. For purposes of comparison, a map of the same area as prepared by the United States Geodetic Service is given also.[8] The last illustration (D) is a map of the Hudson River (1½-mile range) as seen on the PPI screen of a commercial radar receiver on a steamer in the Hudson River opposite Canal Street, New York, N. Y. The New Jersey shoreline is on the left, and the shoreline of Manhattan Island is on the right. The dark strip indicates the river and the white spots in it are water craft of various kinds.

Another important receiver use of the cathode-ray tube is in the panoramic receiver, also known as a Panalyzer, or Panadaptor, which has been used by amateurs for many years. During the war, it proved to be a very valuable piece of communications equipment. The panoramic receiver will be dealt with in greater detail in a later chapter.

The Cathode-Ray Tube as a Generator

The second category of uses mentioned above is not as clearly defined as the first. The equipment as a rule is still experimental in nature. Tubes other than those using luminescent screens, although properly classified as cathode-ray devices, are used as generators of electrical energy of various kinds, and have become commercial realities, as, for example, the Phasitron.[9]

These devices utilize the electron beam for purposes other than to make a screen fluoresce. The beam may be rotated by suitable means and in that way be made to complete alternately electrical circuits with a series of stationary metallic segments, which, in turn, may control other circuits. The beam now acts as an electronic commutator. Because it is almost inertialess, the frequency of commutation can far exceed that of any conceivable mechanical device.

In another use, the beam may excite a resonant cavity, thus producing extremely high frequency oscillations, in a manner not unlike the magnetron. There are applications indicated by patent disclosures listed in the bibliography, but in view of the basic purpose of this book, descriptions of these patents will not be made here.

The Cathode-Ray Tube in Measuring and Testing Equipment

This is the third and most important group of uses, and it is of primary interest to us since its use in the cathode-ray oscilloscope is included. In this capacity, it performs two major roles. The foremost is its function as the display element. Information made available by the unit appears on the screen of the tube. The interpretation of this information involves the proper correlation of the trace with the settings of the auxilliary apparatus, which, together with the tube, comprises the oscilloscope.

Sometimes the complete device is housed in one metal cabinet. At other times, accessory apparatus is used with it, and the complete system then consists of several individual units. Even then, the cathode-ray oscilloscope is a complete individual assembly.

The second major role of the tube is related to its practical purpose. Its purpose is to portray one voltage as a function of another. In general, where the cathode-ray tube is used for instrumentation purposes, electrostatic-deflection methods will be used. This is so because only electrostatic-deflection methods are sufficiently versatile to perform the variety of functions, and reproduce the great variety of waveforms the oscilloscope is called upon to reproduce. Electromagnetic deflection is usually used where the conditions of operation are fixed, as, for example, in television receivers and radar receivers.

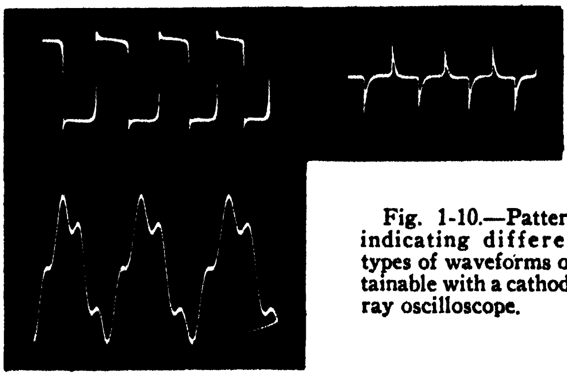


Fig. 1-10.—Patterns indicating different types of waveforms obtainable with a cathode-ray oscilloscope.

Examples of patterns representative of the diversified applications of the cathode-ray oscilloscope appear in Fig. 1-10. All of these picture electrical phenomena. This need not be so, as will be shown later in this chapter, but for the moment, we shall deal with fundamental electrical measurements rather than nonelectrical applications. This raises a very interesting aspect of cathode-ray oscillography, namely, the relationship between the capabilities of the tube and the capabilities of the oscilloscope as a whole.

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[8] Rider, John F., and Rowe, G. C. B., "Radar, What It Is," John F. Rider Publisher, Inc., New York, N. Y., 1946.

[9] Rider, John F., and Uslan, Seymour D., "FM Transmission and Reception," John F. Rider Publisher, Inc., New York, N. Y.; 1948.

The Cathode-Ray Tube and the Oscilloscope

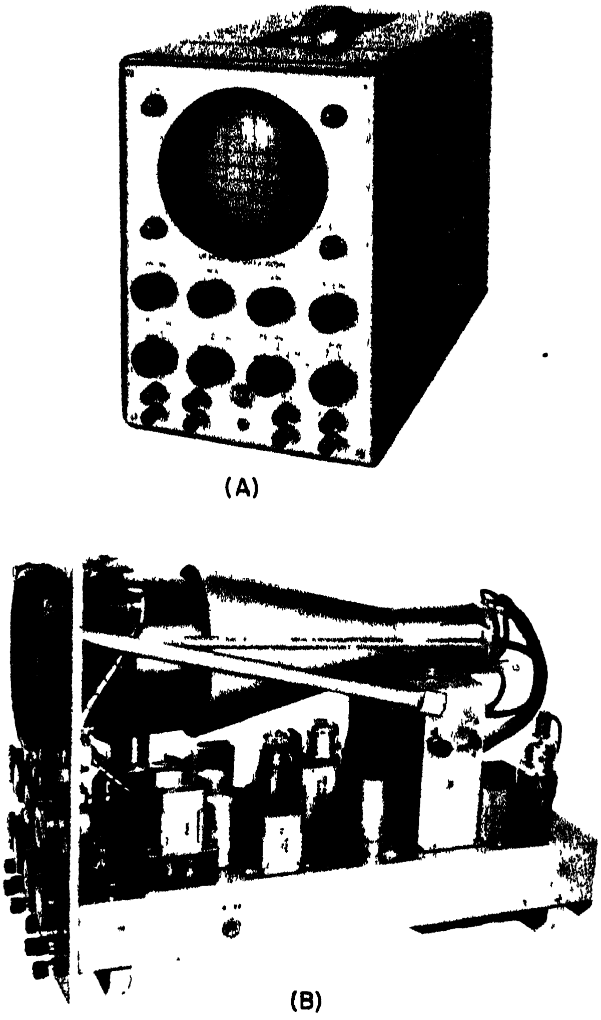
At the time Johnson introduced his low-voltage cathode-ray tube, the cathode-ray oscilloscope consisted of the tube, its operating voltage sources, and the time-base oscillator. Physically, all the parts were breadboard mounted and the tube was completely accessible. As for use, it was recommended for the examination of electrical and magnetic phenomena.

It could do this. The electrostatic field was developed between the surfaces of the deflection plates, and if an electromagnetic field was to be applied for the examination of electromagnetic phenomena, it was a simple operation, since the tube was out in the open. The deflection coils were located at the appropriate points along the neck of the tube. It all was very easy. Such things as amplifiers, used ahead of the deflection system, were yet to come. In fact, the writer detailed the cathode-ray tube, or rather the oscilloscope as described above, which in essence was merely the cathode-ray tube, as a laboratory in itself.[10]

Under the circumstances, it is not surprising that, beginning with its introduction and for many years thereafter, the cathode-ray oscilloscope has consistently been considered as a device suitable for the display of both electrical and magnetic phenomena. However, what was physically true about the first version of this device no longer is true. Modern construction forces a choice between these two faculties. The modern oscilloscope does not afford access to the tube for electromagnetic deflection.

Not only is the electron beam made inaccessible for magnetic deflection, but all stray magnetic fields are excluded; every precaution is taken to shield the electron beam from magnetic influences. This is standard practice in virtually every cathode-ray-tube oscilloscope, even if the measurement is of electromagnetic effects!

As a case in point, and as an example of standard practice, Fig. 1-11B shows the internal construction of the general purpose cathode-ray oscilloscope shown in Fig. 1-11A. Note especially the conically shaped structure which apparently surrounds the cathode-ray tube. It is a *magnetic* shield, usually made of mu-metal, a material of extremely high permeability, offering the best path for magnetic lines of force, and keeping them from affecting the electron beam inside the tube! Another example is illustrated in Fig. 1-12. This device, a **B-H** tracer, is a cathode-ray oscilloscope which is specially designed for the display of the hysteresis curves of ferrous materials. The electron beam within the cathode-ray tube is not deflected by an electromagnetic field; the tube is of the electrostatic-deflection variety. The magnetic properties of the substance being tested are converted into suitable voltages and applied to the electrostatic-deflection plates. The circuits involved are explained and illustrated in the chapter devoted to electrical measurements.



(A)

(B)

Courtesy RCA

Fig 1-11.—(A) General purpose cathode-ray oscilloscope RCA type WO-60C, (B) side view of internal construction.

The foregoing should not be taken to mean that magnetically deflected cathode-ray oscilloscopes do not exist. They are in use, but are not available com-

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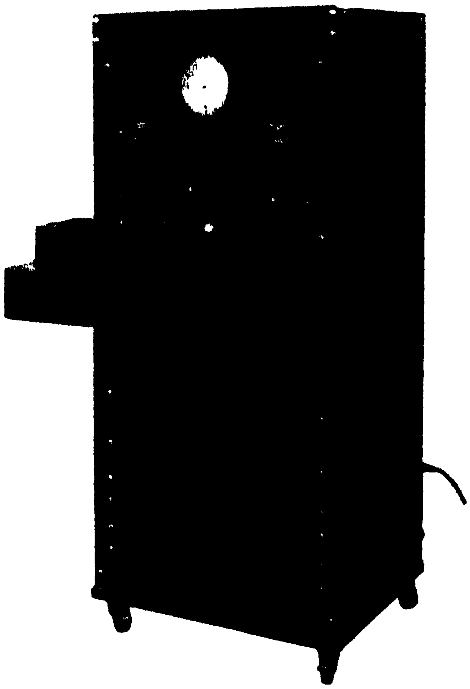
[10] See footnote 6.



Courtesy Brush Development Co.

Fig. 1-12.—B-H curve tracer used for display of hysteresis curves.

mercially. It is, therefore, possible to say that, at this time, the investigation of all phenomena by means of the cathode-ray oscilloscope requires conversion of the phenomena into appropriate deflection voltages.

A Continuous Screen Magnetically Deflected Oscilloscope

It might be of interest to describe briefly a magnetically deflected oscilloscope which might conceivably set a pattern for the future. At the moment, it is a laboratory device and represents an effort to provide continuous display for the study of speech. The limits of the screen in the ordinary oscilloscope require that the electron beam return to its starting point after an excursion across the screen. This is circumvented in this new device by making the tube as a whole rotate so that the screen offers a never-ending recording surface.

The screen is located on the side of the tube in the form of a wide ring. To quote Johnson,[11] "the electron beam is directed initially upward along the axis of the tube in the conventional manner; then it is bent at right angles so that it will strike the screen surface. This is accomplished in the bulbous part of the tube by a steady magnetic field normal to the plane containing the tube axis. Superimposed on the bending of the beam is a variable magnetic deflection governed by a yoke which surrounds the tube at the top of the long neck. By means of this coil the beam is swept up and down across the screen with a linear repetition of about 100 cps. Since the screen is made of the high persistence type of phosphor, a pattern made on it will remain in evidence for several seconds after excitation, which occurs at a point just to the right of the viewing window. The screen persistence is such that the record of the signal appears over about one-sixth of the screen circumference. After a full rotation the intensity of the pattern has decayed to such a low level that a negligible trace of it remains in a new pattern."

In summary, then, the commercial cathode-ray-tube oscilloscope is a voltage-responsive device; we might go so far as to say that its success is based upon a condition very ably stated by Berkley,[12] "in nearly twenty years of oscillography, the Du Mont Laboratories has yet to find a phenomenon which is incapable of conversion into a suitable electrical signal."

COMMERCIAL CATHODE-RAY OSCILLOSCOPES

The cathode-ray oscilloscopes available commercially include a wide variety of instruments. They are divisible into two main categories, however, and we can analyze them better if we treat them in groups. These classifications are:

1. General purpose
2. Special purpose.

Each of these has numerous subdivisions, and an analysis of the manner in which these devices are similar to each other or unlike each other requires a summary of the contents of each classification.

General Purpose Oscilloscopes

This is the group of instruments offered for general use in the electronic industry, for research, maintenance, and production testing, for student use in laboratories, as well as in other branches of science and industry where the phenomenon of interest is electrical in character. It is the kind of equipment which we defined at the very beginning of this chapter. Examples of it are shown in Fig. 1-13. These are but a few of the many different brand names which are available for purchase and are completely described by performance specifications in Chapter 22.

The few examples given in Fig. 1-13 are not representative of the full range of capabilities of the general

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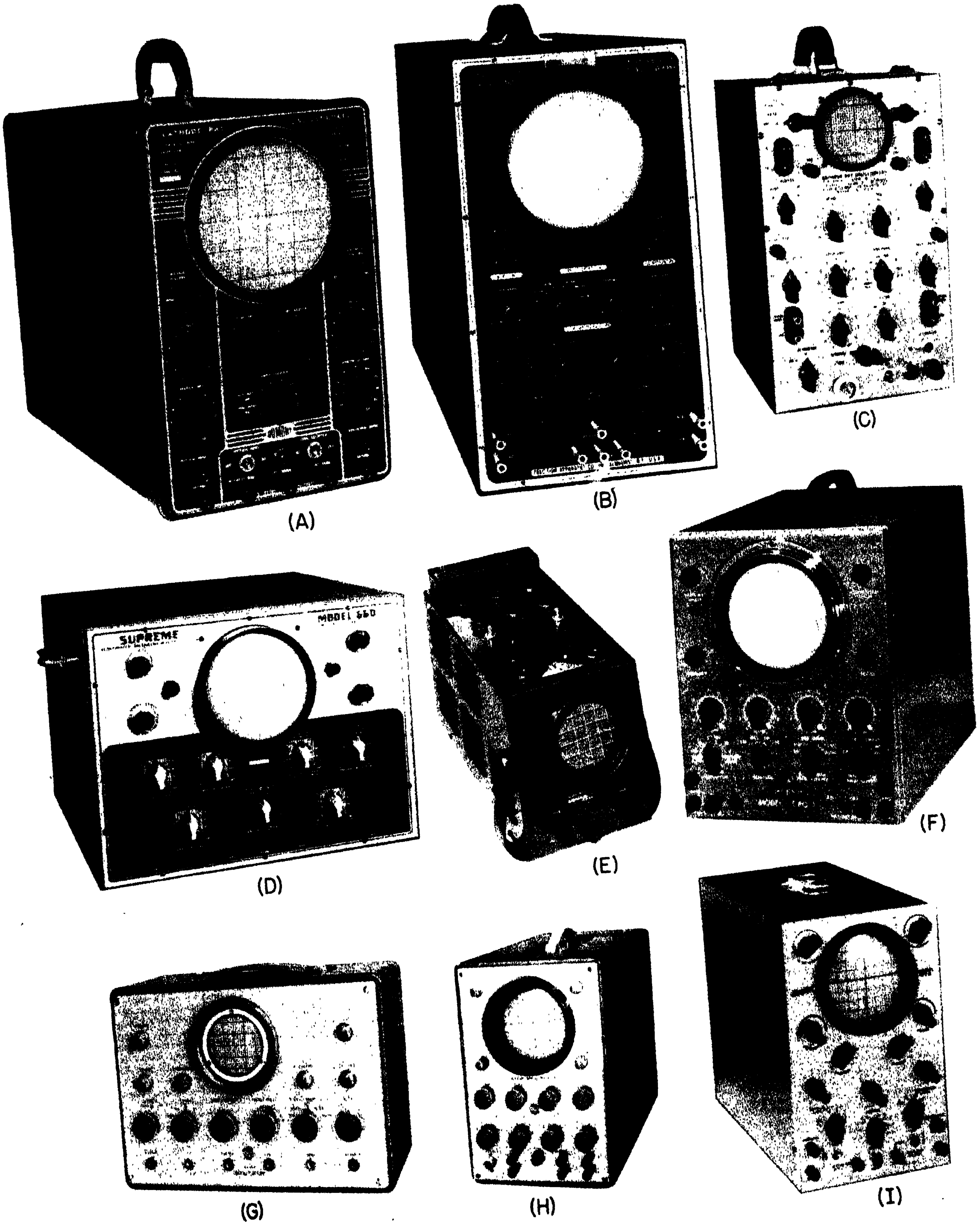
[11] Johnson, J. B., "A movable screen cathode-ray tube," *Bell Lab. Rec.*, vol. 26, p. 219, May, 1948.

[12] Berkley, Carl, "Transducers," *Du Mont Oscillographer*, vol. 12, p. 9, January-March, 1950.



Courtesy of manufacturers listed

Fig. 1-13.—General purpose oscilloscopes; (A) Du Mont 274, (B) Precision ES-500, (C) Du Mont 224A, (D) Supreme 660, (E) Waterman S-14-B, (F) Jackson CRO-1, (G) RCA WO-55A, (H) RCA WO-58A, and (I) Hickok 195B.

purpose category. The general purpose category can be subdivided into smaller groups, each distinctive in the performance specifications associated with the group. We shall show some of these later in connection with the expanded breakdown of the classification. Although somewhat similar in appearance, numerous differences are nevertheless to be found among oscillographic instruments within a single group. At the same time, many points of similarity exist too. Recognition of these points of similarity and dissimilarity is very essential, because it is the basis of equipment classification, as well as the basis of selection of instruments to fill specific needs.

Admittedly, the burden of selection rests on the shoulders of the buyer. The manufacturers of equipment are delighted to furnish whatever advice they can, but it is still the responsibility of the prospective user to understand:

1. The requirements of the phenomenon being investigated.
2. The meaning of performance specifications and how they fit the requirements.
3. The meaning of the pattern which appears on the screen of the tube.

With the exception of the last named, the conditions surrounding selection of equipment can best be expressed by two words—*caveat emptor* (let the buyer beware). This should cause no alarm, however, because of the manufacturer's guidance, this text, the very many periodicals and house-organs which afford information, and, above all, because the appreciation of performance characteristics, operating conditions, and pattern interpretation is not as difficult as it may seem.

Differences Between General Purpose Oscilloscopes

The differences one finds between these devices are numerous. Some are major and some are minor in importance. They are electrical as well as physical, and consequently determine performance, appearance, and even convenience of use. Some appeal to the esthetic sense, whereas others appeal to the pocketbook. Some are desired for a specific purpose and some are intended for general oscillography. All in all there are enough varieties to satisfy the most discriminating and the casual purchaser who is confronted with everyday needs. Let us examine some of these differences.

Physical Differences—The most easily discernible differences are physical. In Fig. 1-13 are shown nine different oscilloscopes. As can be noted in the photographs, some of them are identified as "oscillographs" and some as "oscilloscopes." The names are used synonymously and mean one and the same thing. These are products of well-known manufacturers, but our comments apply just as readily to the products of lesser-known organizations, of which there are many.

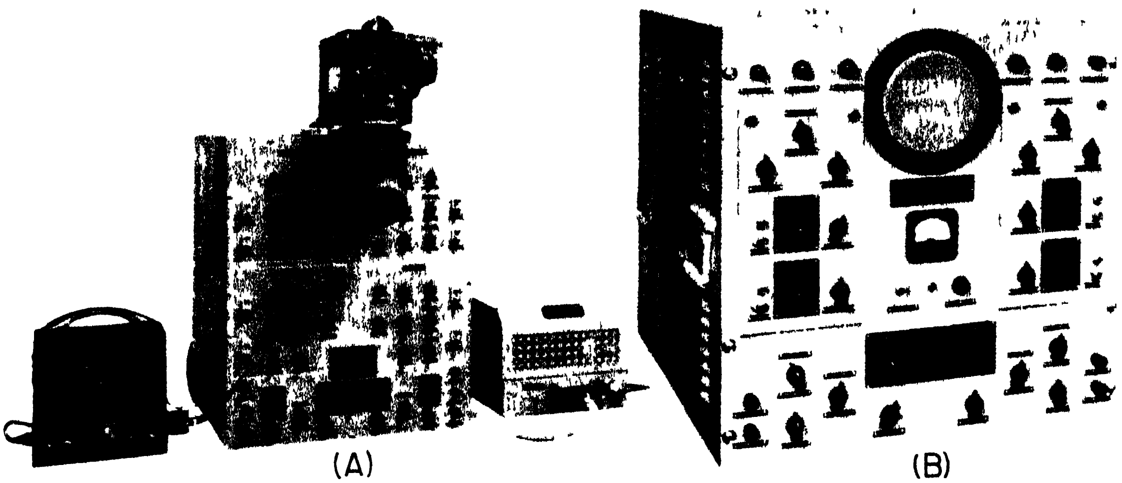
The illustrations do not show the relative dimensions, but it is easy to see that all the instruments are not of the same size. It is also evident that differences in shape, number of controls, probable weight, readability of panel markings, accessibility of panel controls, size of the screen, types of connectors, all are features which distinguish one from the other. What is inside the cabinet is, at this time, an unknown quantity, to be dealt with soon, but for the moment, it is conceivable that a wide variety of physical conditions at the point of use, as well as eye appeal, could result in different choices.

Relative to visible features, a similarity in approach is to be found in the equipment. This is understandable when one appreciates the fact that the various phases of competition exert a tremendous influence. Few manufacturers are willing to make radical changes in the appearance of a device of technical nature after a general pattern has been set by years of production. Devices offered for similar service are made to look as much alike as possible; it tends to minimize confusing influences, yet each producer tries to retain an individuality. Withal, departures from a general pattern of construction and appearance are kept to a minimum.

Occasionally, a manufacturer will come out with a major change in appearance or physical dimension because he feels that the habit of years should be broken. Whether these set new trends or remain distinctly individual, only time will tell. At least there is a variety. One such departure from conventional shape has been in use for a long time, in fact, it is strange to us that it has been alone in its field for even so long.

We make no pretense, at this time, of discussing the relative merits of these differences, although it would be well to mention that they are not as unimportant as one might imagine. Physical features can have a great bearing on the convenience of application. Frequently, the conditions of use are such that correction of a certain form of construction may be necessary. We are speaking at the moment about the readability of the trace, and the panel or dial markings under different conditions of lighting. This may seem trivial, but it is not.

Weight and size are important if a unit must be moved from place to place. Good electrical contacts at connecting points are imperative. Proper panel markings for convenient use are important. These are



Courtesy Electronic Tube Corp

Fig. 1-14—General purpose oscilloscopes utilizing multigun tubes; (A) four-channel with a photographic attachment and (B) two-channel.

practical points of difference. The evaluation of these is a matter of individual taste.

While on the subject of physical differences, we might mention the general purpose oscilloscope which utilizes a multigun tube. Two such are illustrated in Fig. 1-14. Fig. 1-14A is a four-channel oscilloscope with a photographic attachment covering the screen. Fig. 1-14B is a two-channel instrument. Such tubes contain several electron beams which may be operated independently of each other so as to portray two or more distinct phenomena at the same time. The front panel of such a device will bear a resemblance to the ordinary single-beam device, although it will appear more complex by virtue of the duplication of the controls, since each beam is a part of a complete oscilloscope tube system. As far as the tube screen is concerned, only one is used, just as in the case of the conventional tube. The visible evidence of the kind of tube used lies in the duplication of controls rather than in any feature of the tube screen.

Electrical Differences—Everything inside the oscilloscope contributes to its performance; if not to the actual capability of the device, then to the manner in which it fulfills its performance specifications. Such details as the quality of the components, the tolerances employed in manufacture, the general workmanship, and the accuracy of calibration are very important. Some become evident upon inspection, others only during use.

Vertical-Amplifier Bandwidth

A good deal of the utility of a cathode-ray oscilloscope lies in its ability to display voltages of different frequencies. In fact the range of signal frequencies an oscilloscope will pass is one of its prominent performance specifications, if not the most important, and has a great bearing upon its utility in different fields of research. Of the many ways of classifying oscilloscopes, the most effective and most descriptive is the one using the *vertical-amplifier bandwidth*. It expresses the low-frequency limit and the high-frequency limit of the pass-band of the amplifier which feeds the signal to be investigated to the vertical-deflection system of the cathode-ray tube. The dominant role of the bandwidth can be appreciated by considering the simple statement: a 5-Mc signal cannot be passed through an amplifier which has an upper-frequency limit of 2 Mc; or a 1-cycle signal cannot be observed if the lower limit of the amplifier is 10 cycles.

If an oscilloscope is designed for a top frequency of 1 Mc, it is immediately limited in application to those operations which do not demand higher frequencies. In turn if an application demands response at d-c (or zero) frequency, the oscilloscope must be of the d-c variety. If it responds down to a fraction of a cycle, it is not good enough, because there is a wide gulf between d.c. and a fraction of a cycle. These references to frequency have broad implications. They do not refer only to sine waves, but have a great bearing on the quality of reproduction of nonsinusoidal waves, especially square and rectangular pulses. The broader the frequency range, especially on the high end, the more faithful the reproduction of such waves. The same is true at the low end of the frequency pass-band. The lower the low-frequency limit, the lower is the fundamental frequency of nonsinusoidal waves which may be displayed by the device. These details are very pertinent to modern oscillography, because of the increased use of nonsinusoidal waves.

The demands of science and industry and other users of the oscilloscopes are met by the manufacturers by varied equipment designs. Modern oscillography embraces all frequencies between zero and about 10,000 Mc, but available devices have a top limit of about 100 Mc.

Manufacturers have refrained from extensively standardizing the electrical characteristics of oscilloscopes, because every need is different. They differ widely, and the user can best be served by varied designs. This practice makes available to the user what he requires at the lowest price. If all oscilloscopes were standardized so as to encompass all possible uses, many users would find it difficult to acquire such instruments because the costs would be far in excess of what the need warranted. The widespread use of cathode-ray oscilloscopes would be diminished greatly.

So much of the usefulness of the cathode-ray oscilloscope stems from the frequency range over which it is usable, that it is of interest to note the ranges offered by a cross section of the general purpose instruments on the market. These are shown graphically in Fig. 1-15. No attempt is being made to correlate these ranges with the instruments in Fig. 1-13, because the graph includes units not shown in that group of photographs. Although only eleven instruments are represented in the graph, it is possible to arrive at certain conclusions.

We find that some instruments have identical upper-frequency limits for the vertical-amplifier bandpass, but differ at the low end. In contrast, some give similar response at the low end, but differ at the higher, and a general comparison discloses that vertical-amplifier frequency response may vary at both the upper and lower limits of the bandpass. It is sound design and sound selling to serve a market at the lowest possible cost to the customer. The market is very broad in its needs. The users of oscilloscopes have definite ideas of what they want. By offering a wide variety of designs, each user may pick the one which he feels suits him the best. Some types of oscillographic investigations impose very rigid requirements on the equipment, whereas others are more liberal. One kind of work will be quantitative, another qualitative. Satisfaction of these needs is the responsibility of the varied designs. The instruments which will fill these needs must, of necessity, offer different facilities. Many problems in oscillography remain to be settled. All engineers do not agree upon what constitutes a good reproduction of a square wave. Some are more exacting than others. It all depends upon the individual approach and the kind of information sought in the display. Equipment is made to suit whatever school the user belongs to.

The research engineer demands one category of general purpose equipment, whereas the maintenance man, or the production-line tester, or the student in a laboratory requires another. All instruments may have certain characteristics in common as is indicated by the overlapping lines in Fig. 1-15, but other properties of the equipment, not shown in that graph, yet related to the vertical-amplifier bandwidth, will distinguish one device and its utility from another. These differences are not always reflected in a 1-cps lower limit in one

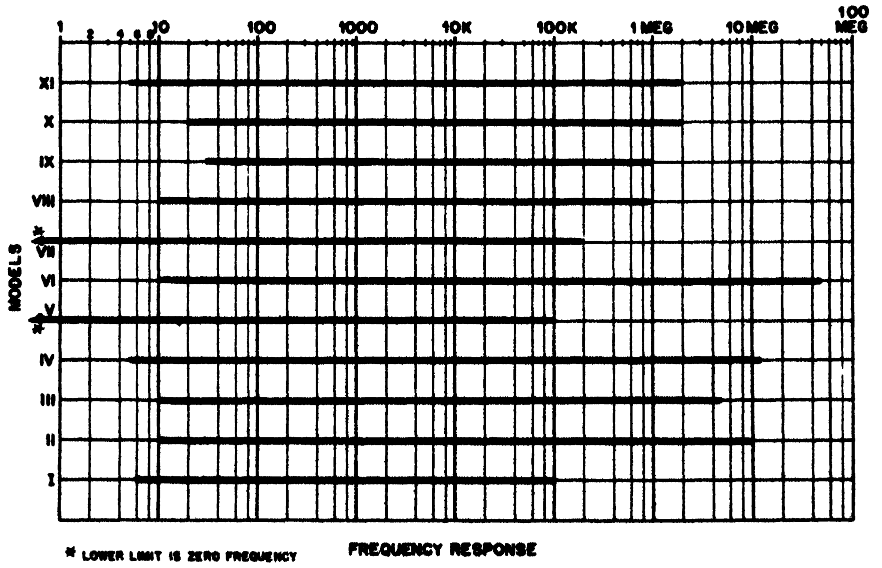


Fig. 1-15.—Graph showing usable frequency range of various oscilloscopes.

device, and a 2- or 3-cps lower limit in another device; or in a 2-Mc top limit in one case and a 1.5-Mc top limit in another. Every limit is a restriction in use but, for all practical purposes, radical departures from the required performance capabilities are necessary in order to prohibit the use of a device in a certain activity.

Such differences exist, as indicated by a 2-Mc upper limit in one case and a 5- or 10-Mc upper limit in another. If the work at hand demands the passing of frequencies up to 10 Mc, then the 2-Mc instrument is useless for that service. But the 2-Mc instrument is usable for general applications of the 10-Mc oscilloscope where the immediate frequency requirements do not exceed 2 Mc. The same is true of instruments with different low-frequency response limits. It is because of such situations that very many different oscilloscopes appear under the heading of general purpose devices, each with different performance specifications.

While on the subject of frequency response, it would be well to emphasize a very important related point. It has a great bearing on the use of the device, although it may not be evident in a simple reference to frequency limits. Although treated in more detail later, this is an appropriate moment to mention the fact that like frequency limits in different oscilloscopes do not mean identical performance and utility. It is a matter of specific amplifier design for qualitative or quantitative measurements. Two amplifiers rated at like bandpass may differ widely in sensitivity. Even with like sensitivity ratings, two amplifiers may be different because the basis of rating the bandwidth may be different. One amplifier may be rated within 1 db and another may be rated within 3 db or even 6 db.

For many applications, this difference in rating may mean little, whereas for others, it may lead to wrong conclusions. From all of this, one can readily see that all oscilloscopes are not alike, although we have mentioned just a few which are associated with one particular section of the general purpose oscilloscope category. It so happens that the section chosen is primary for all uses, and bears a close association with numerous other performance specifications. We do not propose to discuss them all at this time, but a few brief remarks will do no harm to our effort to create some familiarity with other electrical features in which commercial instruments differ. These are indicated very precisely in the equipment specifications.

Pulse Techniques

Investigations in pulse technique constitute a very prominent portion of modern research, and it is not always in connection with television or radar. Many activities utilize pulses to a great extent. The suitability of one oscilloscope of the general purpose variety for this kind of work reflects the difference between that device and some other in the same category. Many observations of pulses are possible on the general purpose oscilloscope, but the more critical study of nonrecurrent, short-duration, transient pulses is possible only with oscilloscopes designed for that function, even though they are included in the broad general purpose classification.

This comes about as a consequence of the circuits and facilities required for the display of these pulse voltages. Although treated in detail in a subsequent chapter, we must mention here that each pulse has a definite frequency content and its proper reproduction on the screen of an oscilloscope demands the passage of all (or nearly all) harmonic components through the vertical amplifier. The absence of certain frequency components can materially distort the shape of the reproduced wave, and thus lessen its value for study purposes. The shorter the duration of the pulse, or the steeper its sides, the greater must be the bandwidth of the amplifier which is passing the pulse voltage. Each oscilloscope has capabilities of this sort, but some have far more than others. The oscilloscope which fits the need is determined by the characteristics of the pulses which must be viewed.

In order to study a quantity which varies with time, regardless of its shape or duration, a time base or "sweep" is required. This, too, is associated very closely with frequency, which may be expressed in units of time. This "sweep" voltage must be variable in frequency so as to accommodate the range of frequencies which is passed by the vertical amplifier. The relationship between the frequency of the time base and the quantity being studied depends upon a number of conditions. The time-base system must give time intervals which are shorter (or longer) than the period of the quantity being studied. This enables one or more cycles (or a fraction of a cycle) of the unknown to be shown on the screen. Normally, the fewer the number of such cycles, the easier studying the phenomenon becomes, but because of conditions associated with the manner of operation of the usual time-base oscillator, at least three cycles of the recurrent voltage should be shown on the screen. This means that the frequency of the time-base oscillator should be variable between one-third of the lower-frequency limit of the vertical amplifier, and one-third of the upper-frequency limit of the vertical amplifier.

If the vertical-amplifier bandwidth is from 3 cps to 1 Mc, the time-base oscillator should have a frequency range variable between 1 cps and approximately 300,000 cps. This has not been a general practice, although it is becoming one. It must be stated, however, that several other factors are related to the operation of the oscilloscope at frequencies below 2 or 3 cps. We are referring to the persistence rating of the screen. Further details of this phenomenon are given in later chapters.*

In addition to the above, another important operating characteristic is associated with the time-base system. This is the kind of time-base voltage required for the display of nonrecurrent transients. The observation of recurrent waves demands a repetitive sweep which repeats itself until stopped. The observation of pulses which only repeat themselves at infrequent intervals, or which last for only a very short time, demands a type of sweep or time-base voltage known as a "single sweep" or a "triggered sweep." With such a sweep, the horizontal displacement of the beam across the screen corresponding to time takes place just once and it is timed in such a way as to coincide with the existence of the pulse. The name "triggered sweep" indicates that the signal being observed sets the oscilloscope sweep into action. Sometimes the reverse is true, that is, the time-base voltage is arranged to sweep the beam across the tube and also start the equipment from which the quantity being studied arises.

Such single sweeps or triggered sweeps are singular features of some general purpose oscilloscopes, and those which contain it are set apart from the remainder, being suitable for special measurements in addition to the remainder of their general uses. It might seem that instruments which have such special facilities should be removed from the general purpose classification. They receive recognition and even special identification, as will be shown later, but are left in the general purpose category because many of them are useful for numerous other everyday functions. Those oscilloscopes which have the means of observing transients and short-duration pulses, but are lacking the facilities for everyday work, and there are such, are classified in the special purpose category and do not appear among the general purpose instruments.

Still another electrical feature found in some general purpose oscilloscopes is the means for determining time intervals between instants of an action, or for determining the duration of an action. Such systems are known as timing markers. Two types are available, but they need not be discussed here. It is sufficient, for the present, to say that they represent another point of difference which sets one general purpose oscilloscope apart from another.

Before concluding this portion of the subject, it might be well to mention the fact that the horizontal-deflection amplifier also is a part of the oscilloscope. This is the amplifier which controls the time-base or sweep voltage that is fed to the horizontal-deflection plates. It too has a frequency bandwidth rating, which must, of necessity, be able to accommodate the frequency components present in the time-base voltage and so permit that voltage to retain its complete harmonic content when applied to the horizontal-deflection plates. The same amplifier is usable for other purposes, too, but that is a separate subject.

Types of General Purpose Oscilloscopes

A fundamental characteristic separating types of general purpose oscilloscopes is the nature of the signal voltage acceptable to the vertical amplifier, that is, whether it is d.c. and a.c., or a.c. only. This is indicated by the low-frequency limit of that amplifier. If it is rated at zero frequency, it will amplify a-c and d-c voltage signals with equal facility, but if it has a frequency limit greater than zero, it is suitable for a-c signal voltages only. The difference arises from the design of the coupling system between the amplifier tubes. If they are direct-coupled, that is, if the coupling capacitor is omitted, the amplifier will accept d-c signals and also a-c signals. But if the coupling system employs conventional *R-C* coupling, the amplifier will accept only a-c signals for amplification. The d-c amplifier will transmit a constant voltage as well as a varying voltage, whereas the a-c amplifier, by virtue of the presence of the coupling capacitor, demands a voltage variation.

Designing the vertical amplifier so that it accepts zero-frequency signals has a great effect on its bandpass. It narrows the bandpass so much that it is permissible to make the statement that d-c amplifiers invariably are low-band amplifiers. At the present time, the oscilloscopes which offer this facility usually have a top frequency limit of about 200 kc, frequently less. If the shape of the response curve is not too important, that is, if a reduction in amplification of more than 6 db from the "flat" portion of the curve is acceptable, one might say that response up to 300 kc is available in some of these units, although the over-all utility of this extension in the frequency range is very slight since the primary purpose of the design is improved response at the low end of the frequency range.

Amplifiers classified as a-c units have a very wide range of bandwidths extending from some small value such as 2 or 3 cps to as high as 100 Mc. This range is not, as a rule, available in a single unit. Usually, the higher the upper-frequency limit, the more restricted is the lower limit, but this is not a standard practice, as indicated in Fig. 1-15.

Identification of Bandwidth Ratings—In view of the different bandwidth ratings of the vertical amplifiers in oscilloscopes, it seems appropriate to establish identifying terms which will enable one to readily recognize the classification of the device. As we analyze the products on the market, we visualize three categories. Some individuals may not agree with our choice of names, or with the basis we use for establishing the types. The final decision is in the hands of the industry; what is offered here may be considered as a beginning.

As we see it, three band-pass groups exist. These are:

1. Wide band
2. Medium band
3. Low band.

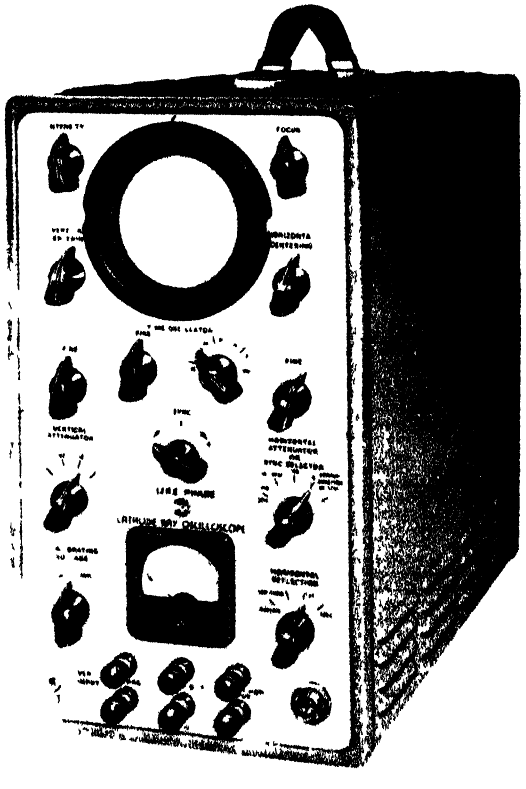
By *wide band* we mean vertical amplifiers with upper-frequency limits between 3 and 100 Mc. The frequency limit at the low end of the pass-band is being ignored temporarily, at least in a-c amplifiers, because the primary interest in the majority of the applications is in the upper-frequency limit. Thus any oscilloscope which contains a vertical amplifier with a frequency response rating of 3 Mc or higher, we consider to be a wide-band device. We might subdivide the 3- to 100-Mc range into such classes as "extra-wide band" or "super-wide band," but these do not seem fitting, since it is reasonable to assume that the present-day commercial limit of 100 Mc will be exceeded in the not-too-distant future.

By *medium band* we mean vertical amplifiers which will accept frequencies up to 3 Mc. This top limit is an arbitrary choice, as is its use as the base frequency for the wide-band amplifier.

The *low band* extends up to 500 kc. This too is an arbitrary choice and one might say that the limit should be lower or possibly higher. Let the future decide that. In the meantime, the manufacturers of oscilloscopes or the committee responsible for the standardization of oscilloscope nomenclature has something to masticate.

Facilities Associated with Classifications of Oscilloscopes—Continuing with the breakdown of the general purpose oscilloscope, each category of instruments within a broad classification offers one or more facilities that are not available in the other categories. All of these facilities may not be duplicated exactly in all examples of that category, but a general similarity exists. However, because of the fact that each manufacturer has his own ideas of what he can include in a piece of equipment sold for a certain price, the existence of the facilities may be assumed, but verification is necessary. Placing a number of different products within a certain category because their vertical-amplifier bandwidths fall within the limits of that category is not necessarily positive proof that the device contains the same circuits, or even similar circuits and facilities as another with the same vertical-amplifier bandwidth ratings. More than likely, similar service may be performed with both devices, but the final answer lies in the individual performance specifications.

The general practice in the design of cathode-ray oscilloscopes is to include certain features in a wide-band instrument. Obviously, the wider the pass-band, the greater may be the number of these features, although a unit with a reduced bandwidth rating within the same category, or in another group, may contain similar features, but not necessarily to the same extent. On the other hand, a device rated at a much lower bandwidth may also have such features, because they may be required in connection with a special activity for which the device is suitable within the lower-frequency range. Strict adherence to the meaning of these categories may prove misleading. That is the reason for this reference. Generalizing as we have done



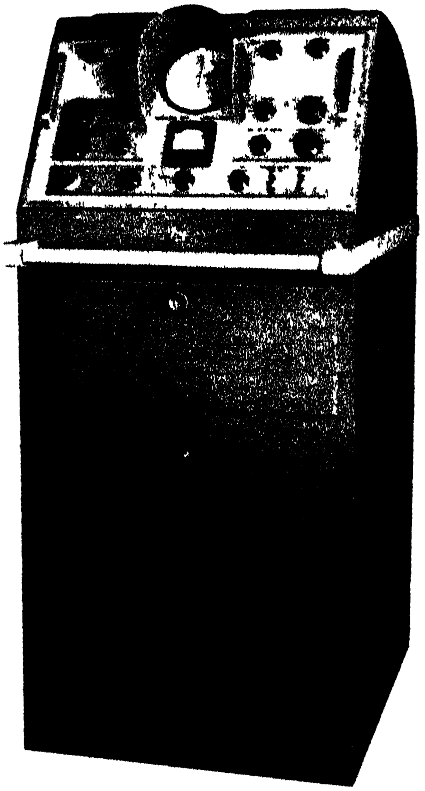
Courtesy RCA

Fig. 1-16.—Wide-band oscilloscope containing a vertical amplitude calibrator.

is a satisfactory means of describing a situation, but it is necessary to drop a word of caution about the contradictions which exist. They are unavoidable because of the great overlapping in design capabilities caused by the overlapping in requirements.

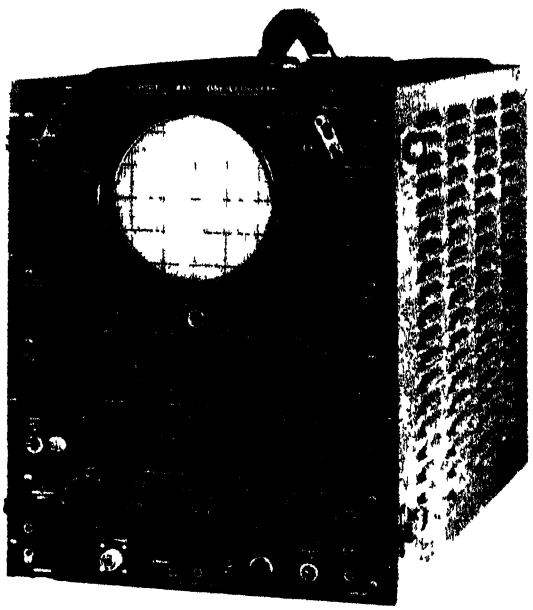
The group of photographs in Fig. 1-13 contains examples of the three categories of general purpose oscilloscopes, namely, wide band, medium band, and low band. In the first named, which seems to be the most prominent category, the examples given have numerous features in common, but also several features which are unique. Among these are the rated bandwidths and the presence of the *Z*-axis amplifier for intensity modulation. In contrast is the instrument shown in Fig. 1-16. This is a wide-band device with a reasonably high bandwidth rating, containing a vertical amplitude calibrator which is not inherent to the units in the group picture. On the other hand, the instrument in Fig. 1-16 does not contain the *Z*-axis amplifier for intensity modulation on the face of the panel. No attempt is being made to compare the utility of these instruments by means of these simple references. It cannot be done this way. The few performance specifications we may mention convey some information, naturally, but it is not the entire story. All the performance specifications must be examined. The brief references being made here are intended to convey the fact that all devices which are grouped together into a classification are not necessarily alike.

To demonstrate this still further, Figs. 1-17, 1-18, 1-19, and 1-20 are examples of four more wide-band general purpose oscilloscopes. The first of these is re-



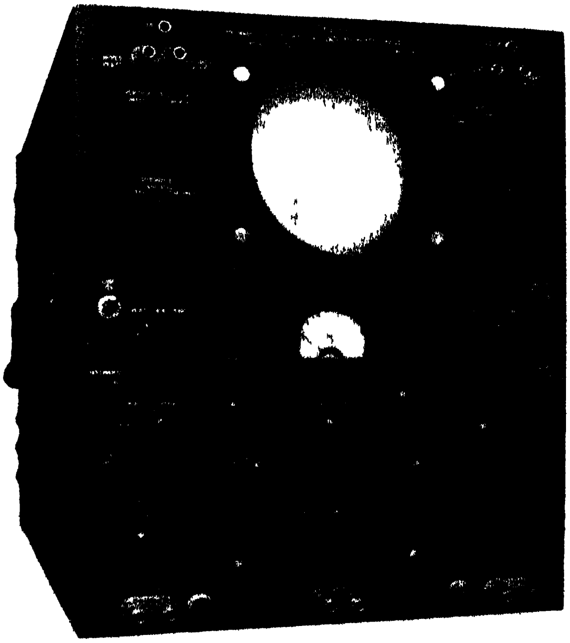
Courtesy RCA

Fig. 1-17.—Wide-band oscilloscope suitable for laboratory use.



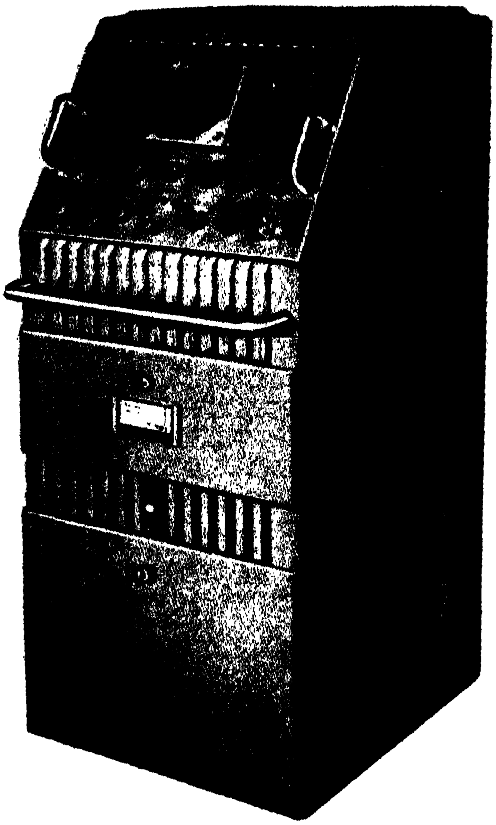
Courtesy Du Mont Labs

Fig 1-18.—Wide-band oscilloscope, Du Mont type 248.



Courtesy Browning Labs.

Fig. 1-19.—Wide-band oscillosynchroscope, Browning type OL-15B.

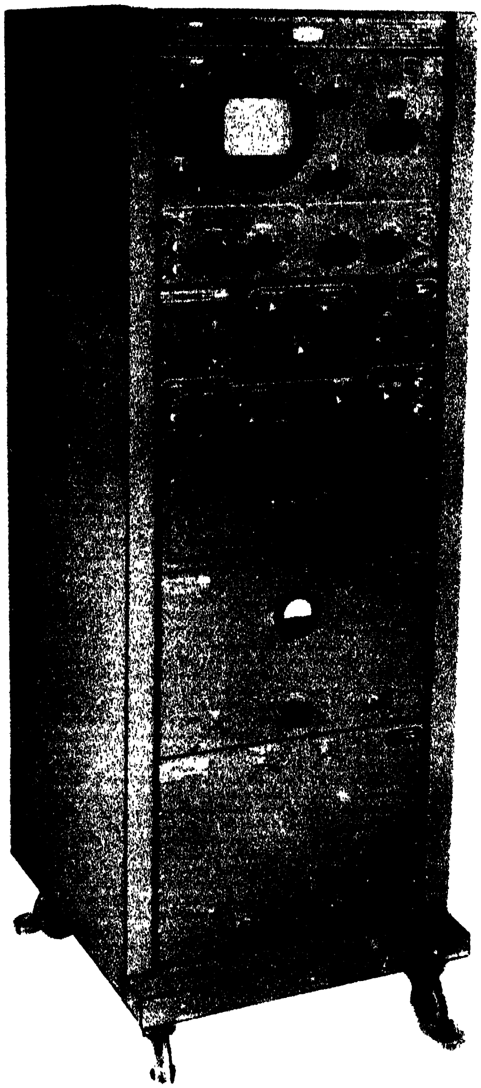


Courtesy Federal Telecommunication Labs.

Fig. 1-20.—A 50-Mc wide-band video oscilloscope.

ferred to in the manufacturer's literature as a television research oscilloscope because of its characteristics, although the facilities it contains are useful for many other activities. The devices shown in Figs. 1-18, 1-19, and 1-20 also are general purpose oscilloscopes, containing still more facilities than are available in Fig. 1-17 for the individual who demands them.

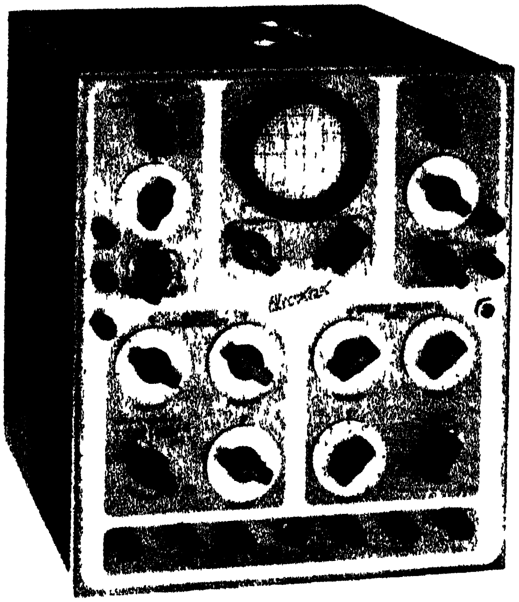
In the majority of instances, these extra facilities are electrical in nature, like those mentioned earlier. They are also physical in nature, as for example, in Figs. 1-16 and 1-17. These are two products made by the same manufacturer. While there are electrical differences between the two units, the fundamental physical difference is found in the mobility of one as compared with the other. One is a portable device intended for use on a bench, whereas the other is mounted on casters so that it may serve any number of individuals in a laboratory. Because of the implications of illustrations, we desire to say that the range of available instruments is not limited to those which we have illustrated. Many other products of different manufacturers exist on the market, but it is impossible to show pictures of all of them. Moreover the references to differences between them must be general. Any attempt to be specific would result in a duplication of the performance specifications contained in Chapter 22.



Courtesy Du Mont Labs.

Fig. 1-21.—Special general purpose oscilloscope for television research or broadcast station use.

Special General Purpose Oscilloscopes—Sometimes a general purpose oscilloscope is contained within a housing that also contains supplementary equipment with which special functions may be performed. In a way, such a device is a special purpose oscilloscope, but in view of the utility of the oscilloscope for performing general functions, we have elected to include them under a special heading of the general purpose group. The reason for this is their association with the electronic industry, in contrast with the special purpose oscilloscopes intended for nonelectronic applications.



Courtesy Hickok Electrical Instruments Co

Fig. 1-22.—Oscilloscope containing special signal source for alignment of any system demanding an f-m signal.

One example of a special general purpose oscilloscope is shown in Fig. 1-21. This device is designed for use in television laboratories or at television broadcast stations for checking the characteristics of television signals. Another example of a device within the same category but intended for an entirely different purpose is shown in Fig. 1-22. This is a combination device containing a general purpose oscilloscope and the special signal source required for alignment operations on a-m, f-m, and television receivers, or any type of system which demands a frequency-modulated signal. In general it is an alignment device, but when desired, the two major sections, that is, the oscilloscope or the sweep generator, may be utilized independently of each other.

The equivalent of such a combination unit is quite frequently formed by the use of two individual units, a sweep generator and a general purpose oscilloscope. In fact, these are more commonly in use than the combination instrument. That this should be the case is not surprising, since then each device may be used for applications which are unrelated to each other. On the other hand, the combination device has its advantages too in that the integration of the two systems affords great convenience. Again it is a matter of individual preference and needs.

Special Applications of General Purpose Oscilloscopes

It is very important to understand that general purpose oscilloscopes may be employed to perform many special functions. If the correct type of instrument is available for the conversion of a phenomenon of one kind into a suitable voltage, the general purpose unit can be applied to many uses, thus becoming a special purpose oscilloscope without in any way impairing its general purpose utility. For example, a vibration pick-up, of which there are many kinds to suit many different purposes, can adapt a cathode-ray oscilloscope for a wide variety of vibration measurements and studies. It may be such a simple thing as checking the vibration of an engine bed, turntable support, radio cabinets, or the like, or it may be applied to such massive objects as bridge platforms and columns in buildings. Mechanical motion of all sorts, such as that in camera shutters, relays, rotating contacts, the action of a spring, or weaving machinery may be observed. Variations in pressure or strain are commonplace measurements made with such devices. Whether they are made with general purpose oscilloscopes operated in conjunction with a transducer or if they are specially designed units is determined by the nature of the work, but, as we remarked earlier, the simpler of these can be accomplished with the general purpose instrument.

Numerous electromagnetic measurements and studies are being carried on today using the general purpose oscilloscope. The interrelation between magnetic and electrical phenomena makes such operations relatively simple. Motion of a conductor in a magnetic field will result in a voltage which can very readily be correlated with the original motion. Depending entirely on the transducing device, phenomena such as light and heat come within the province of the general purpose device. Even the measurement of X rays and other kinds of rays are within its capabilities by using the Geiger counter as the transducer.

These illustrations are but a few of the great number of applications, very many of which are shown and described in the industrial and scientific applications chapters. The educational applications of the general purpose oscilloscope also are very numerous and these too are described later in this text.

Summary of General Purpose Oscilloscope Facilities

If we select the significant features in the different classifications of general purpose oscilloscopes as a basis of comparison, neglecting for the moment all of the common characteristics, it is possible to develop a chart which will show the correlation between these features and the categories. Such a tabulation follows.

- General Purpose Oscilloscopes
  - Alternating Current
    - Wide band (3 Mc and higher)
      - triggered sweep
      - *Z*-axis amplifier
      - timing markers*
      - specific sweep speeds
      - repetitive sweep (variable frequency)
      - calibrator*
    - Medium band (500 kc to 3 Mc)
      - triggered sweep*
      - *Z*-axis amplifier*
      - repetitive sweep (variable frequency)
      - calibrator*
    - Low band (up to 500 kc)
      - triggered sweep*
      - repetitive sweep (variable frequency)
  - Direct Current
    - Low band (d-c to 300 kc)
      - repetitive sweep (variable frequency)
      - triggered sweep*
      - timing markers*

*available in some instruments.

Other significant specifications may have been selected in addition to those shown, but the list above is representative of what one might expect when the device is identified in terms of the vertical-amplifier bandwidth.

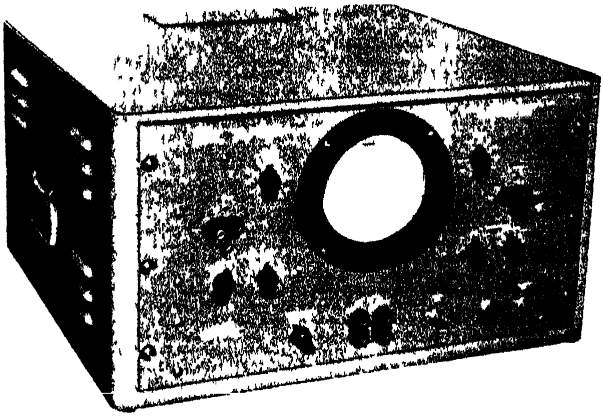
The Special Purpose Oscilloscope

Again we encounter a wide variety of instruments. Special purpose oscilloscopes are of singular character in that each device is designed to perform just one role. It may be designed to serve an industry by being calibrated in terms of the units used in that industry. It may be a device which is not a complete oscilloscope, but instead is the cathode-ray tube and its power supply only, thus allowing it to be used as simply a cathode-ray indicator or display. As we remarked earlier, many varied services may be performed by the oscilloscope so that any instrument which fits a field of activity especially well might be viewed as a special purpose device.

In view of the myriads of measurements which are possible with the equipment, it is hoped that a few examples will suffice in this introductory chapter. All of those given are described in terms of the basic phenomena and not the means employed to accomplish the display. The story of the transducer, the phenomenon-converting device, is a long one, and is discussed in the chapter devoted to scientific and industrial applications. The fact that an oscilloscope may be designated as a special purpose device does not immediately remove it from the electronic or the electrical industry, and place it in one of the nonelectrical classifications. It may be used in all branches of science, industry, and education. It becomes special purpose because it is designed to best fit a certain need, and its utility as a general purpose device is kept to a minimum if it exists at all.

Cathode-Ray-Tube Indicator

The simplest example of a special purpose oscilloscope, although in reality it is not an oscilloscope, but rather just a cathode-ray-tube indicator, is shown in Fig. 1-23. It consists of a cathode-ray tube, its power supply, and the required beam-positioning controls. All amplifiers and time-base systems are missing, it being assumed that whatever systems are required to make the device perform its function (exclusive of the power supply) will be furnished externally. As shown, the tube may as readily be the balance indicator in a bridge as the modulation indicator in a transmitter. As a matter of fact, this and several other brands of like devices are offered in this manner to fit any application which requires a cathode-ray display tube using an electrostatically deflected cathode-ray tube.

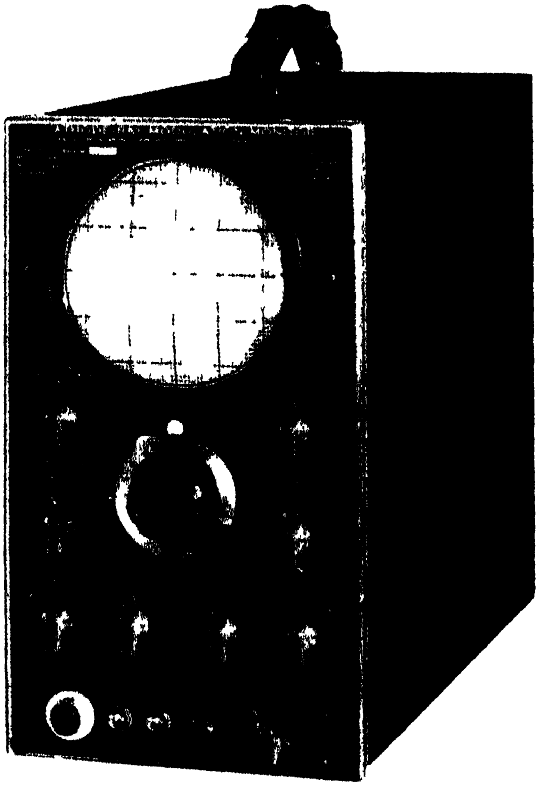


Courtesy Du Mont Labs

Fig 1-23.--Unit including only cathode-ray tube, power supply, and beam-positioning circuits

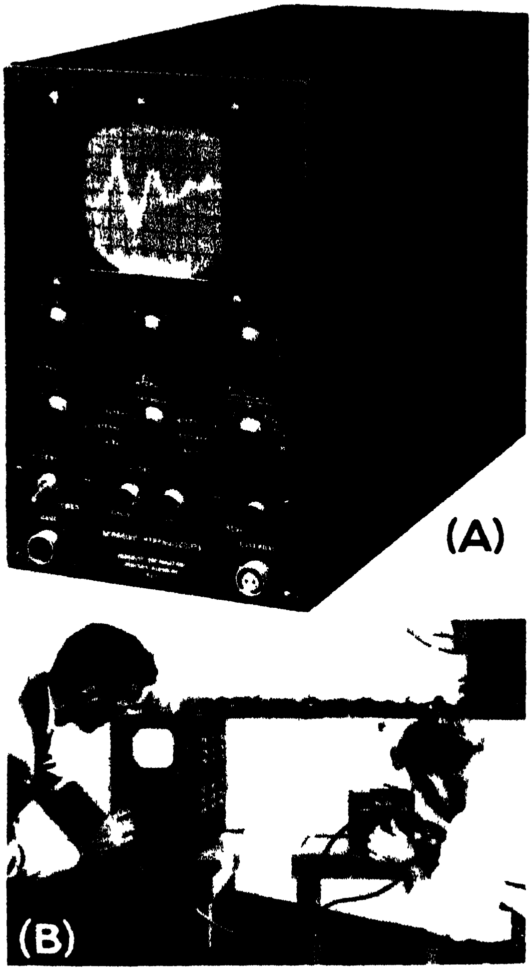
Modulation Monitor

An elaboration of the fundamental device appears in Fig. 1-24. This unit is a special purpose oscilloscope designed specifically for monitoring the output of transmitters and is called a cathode-ray modulation monitor. It is more elaborate than the basic indicator in that it contains the means for showing the modulated wave envelope as well as the trapezoidal modulation pattern. It contains a variable tuned circuit tunable over a fre-



Courtesy Du Mont Labs

Fig. 1-24 —Cathode-ray unit used to monitor the output of a transmitter.



Courtesy Aeroquip Corp

Fig. 1-25.—(A) Measurement of static and transient pressures using an Aeroquip hydrauliscope, (B) device being used to check out the hydraulic system of the Martin 2-O-2.

quency range of from 0 4 to 40 Mc and a variable-frequency time-base or sweep system covering a range of from about 15 cps to above 30,000 cps. The tuned circuit feeds the carrier signal to the device for either wave-envelope or trapezoidal-pattern display.

B-H *Curve Tracer*

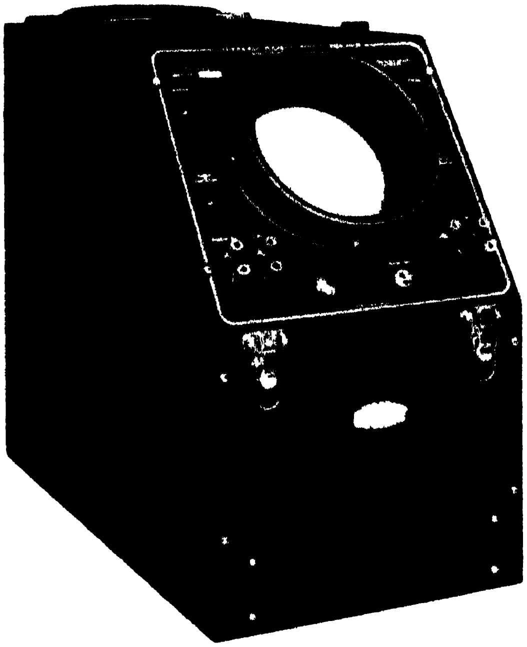
The **B-H** curve tracer shown in Fig. 1-12 is another example of a special purpose oscilloscope, being intended specifically for the investigation of magnetic materials. It contains all of the elements of the fundamental cathode-ray oscilloscope, but these are arranged in such fashion as to be unavailable for general purpose work, and yet afford the utmost in convenience for the special purpose for which it was designed. It is the first of the instruments we have shown which contains a transducer, by means of which a phenomenon of one kind, in this case magnetic, is converted into voltage for application to the deflection system inside the cathode-ray tube.

Aircraft Measurements

The application of the cathode-ray oscilloscope to the measurement of static and transient pressures in the hydraulic system of an airplane is shown in Figs. 1-25A and B. This is a special purpose instrument used in conjunction with a transducer which transforms the pressure levels into equivalent voltages and applies them to the deflection system of the electrostatic type of cathode-ray tube. The essential elements of a conventional cathode-ray oscilloscope are contained within the main cabinet, but in order to afford utmost convenience in the application for which the equipment was designed, only those controls are available as are essential to the attainment of its objective.

Polar Co-ordinate Indicator

Still another example of a special purpose oscilloscope is shown in Fig. 1-26. This device, called a polar co-ordinate indicator, is designed for the study of rotating machinery of all kinds, and ideally suits the needs of the automotive industry for the study of distributor action, motor and generator brushes, and the like. Further details about this device and the Hydrau-



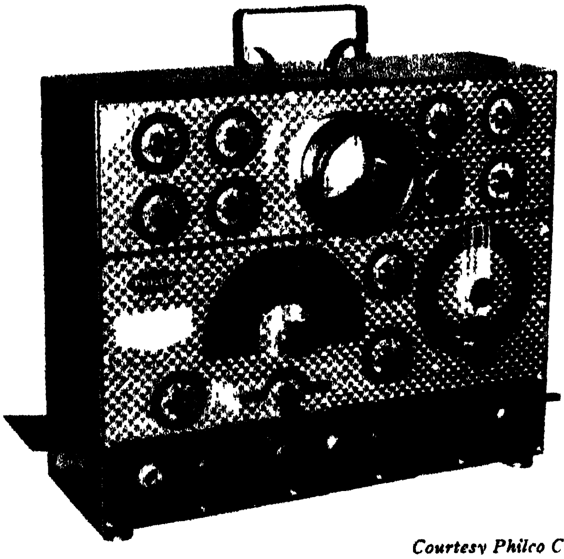
Courtesy Du Mont Labs

Fig. 1-26.—Cathode-ray polar co-ordinator designed for the study of rotating machinery.

liscope will be found in the chapter devoted to scientific and industrial applications.

Servicing Unit

An example of a special purpose oscilloscope of the simpler variety and intended for radio and television applications in the maintenance and manufacturing fields is shown in Fig. 1-27. It contains a variety of signal sources for all kinds of alignment operations and a cathode-ray oscilloscope for display purposes. All are housed within a single cabinet. The oscilloscope portion contains the necessary beam-positioning control, a vertical amplifier, and a single-frequency sine-wave sweep necessary for alignment purposes. Of course, the device as a whole also is useful in those operations which require the services of various signal sources in connection with maintenance and production tests and measurements.



Courtesy Philco Corp.

Fig. 1-27.—Philco 7008 f-m and television alignment instrument which contains a basic oscilloscope unit.

Concluding Remarks

So much for the general introduction to the different kinds of oscilloscopes. We are aware that some of the statements made here may not be as meaningful as one may desire, but this is due solely to the fact that so much more follows and we are attempting to advance in small stages. This introduction is intended to merely highlight the salient points. Because of the ramifications of oscillography, boundaries must be imposed on the breadth of any discussion because some features demand an appreciation of the underlying principles. Since these are given later in the text, we felt it advisable to refrain from even mentioning certain applications, and above all, from describing fully the potentialities of different instruments.

The examples chosen for discussion and illustration do not imply any preference, nor do the omissions of certain manufacturers' products reflect in any way upon their products. It is just that space prevents illustrating every available device. Some manufacturers are more prominently represented than others, but only because they produce a greater variety of instruments. The absence of some oscilloscopes from the list of performance specifications in Chapter 22 is due either to the unavailability of the information or to the fact that the instruments made their appearance after the book was on the press. Approximately 80 different models of oscilloscopes are described later, and this is a very substantial coverage of the commercial market. We repeat here what we have said before, namely, that comparison of instruments can best be accomplished by an analysis of the performance specifications, and now we add that after those findings are made, the reputation of the manufacturer and what he puts into his product must be given cognizance also.

The omission of a reference to some special application which may come to the mind of the reader is not a sign that it has been either neglected or forgotten. More than likely, it will be found somewhere else in this book, for every attempt has been made to cover the field of oscillography as completely as possible, bearing in mind that the content of this book is devoted to the application of the cathode-ray-tube oscilloscope as it relates to measurements, research, and education.

CHAPTER 2

THE PRINCIPLES OF ELECTROSTATIC DEFLECTION AND FOCUSING

Two basic phenomena pertinent to utility of the cathode-ray tube are focusing and deflection. The first of these is control of the electrons issuing from the emitter so that they will be formed into a concentrated beam which will produce a sharp, bright trace on the screen of the tube. This action is paramount in every cathode-ray tube regardless of its type or ultimate purpose. It governs the application of the tube as a display, in that a poorly focused electron beam will impair the usefulness of the device as a whole. If the amount of defocusing is extreme, it will prevent the appearance of any sort of defined trace on the screen; instead of a trace, the screen will display a dim glow of light across its entire surface. Obviously, therefore, proper focusing is a very important action.

Focusing has a substantial effect upon the intensity of the trace, because it determines the *area* in which the electrons strike the atoms of the screen material. The smaller this area, the greater is the concentration of energy imparted to the atoms of the screen coating by the impinging electrons. The focusing of the electron beam has its equivalent in the focusing of a light beam by means of lenses so as to concentrate the maximum amount of light on a very small area.

Electrons in motion behave in many respects like light rays, and can be focused by means of electron lenses which are specially shaped *electrostatic or electromagnetic fields.* Just as a light lens can alter the direction of light rays, so can electrostatic and electromagnetic fields alter the direction of movement of electrons. This action is known as *refraction,* or bending. How it is done will be explained in this chapter.

The need for focusing the emitted electrons into a beam arises from the condition that the source of emission releases electrons which are scattered in many forward directions as shown in Fig. 2-1. Only a portion of the electrons emitted eventually reach a single focus, but a satisfactory emitter in an electron gun releases a sufficiently great number of electrons per unit time so that, even if only a relatively limited quantity of these are properly focused, an electron beam of adequate density is obtained.

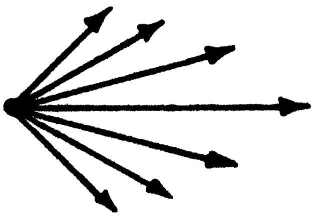


Fig. 2-1. — Paths of electrons after leaving emitting surface of cathode.

Focusing is one of the functions of the electron gun, although focusing of the beam by elements *entirely within* the gun is accomplished only in the *electrostatic deflection* type of cathode-ray tube. In those which are known as the *electromagnetic type,* only a *part* of the focusing action is accomplished by the elements within the gun assembly. This portion is electrostatic even if the tube is of the magnetic kind. The remainder, and by far the more important action, takes place within the physical boundaries of the gun, although not between elements which are an integral part of the gun. The electromagnetic focusing coil is placed in proximity to one of the gun elements, but it is not actually a part of the gun. The relative positions of these elements for both electrostatic and electromagnetic type tubes are shown in Figs. 2-2 and 2-3.

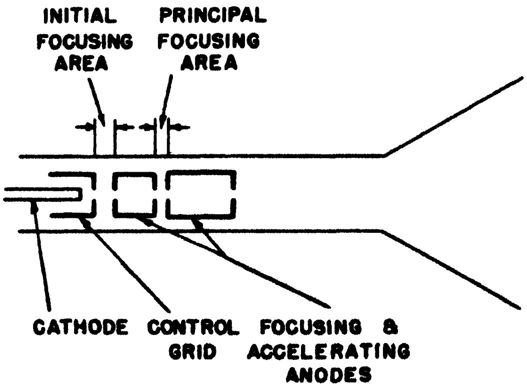


Fig. 2-2.—Relative positions of elements of electron gun within electrostatic cathode-ray tube.

Relative to the system wherein the electron gun elements perform the entire focusing, that is, the electrostatically focused tube, the gun contains the required

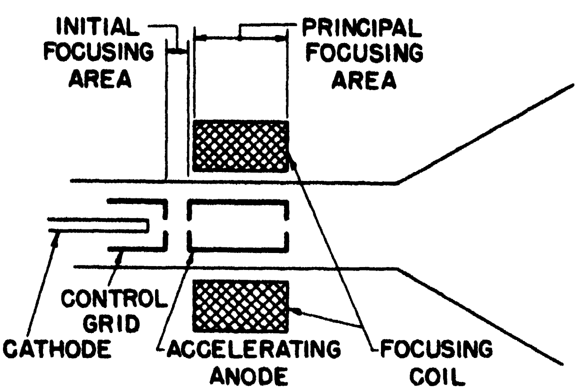


FIG. 2-3.—Relative positions of electron-gun elements and focus coil of electromagnetic cathode-ray tube.

electron lenses, which are specially shaped electrostatic fields formed between electrically charged cylinders arranged in axial symmetry. Although the basic action of electrostatic fields is generally well-known, certain aspects of such fields as employed in electron optics, or as employed for focusing in cathode-ray tubes, are unusual. They differ so much from everyday applications in radio tubes that they deserve more than just casual comment.

In the electromagnetically focused tube, both electrostatic and electromagnetic focusing are employed. This may seem paradoxical in the light of the identification. Actually, such is not the case, because that portion of the electrostatic focusing system which is used in both types of tubes has other functions, too, which are common to all types of cathode-ray tubes. Its contribution to focusing is not its primary function, as will be shown later, which is why the *principal* focusing system remains the basis of identification.

As a general rule, electrostatic focusing is employed in cathode-ray oscillographs and synchroscopes and in small-size television picture displays. It is also used in tubes utilized in other types of equipment, such as the various kinds of radar apparatus. Electromagnetic focusing is generally employed in television receivers which develop large-size pictures and in the larger-size tubes used in radar and similar apparatus; seldom, if ever, is it used in cathode-ray oscillographs for laboratory and maintenance applications.

In order to explain how electrostatic and electromagnetic focusing and deflection are accomplished, we shall review in brief the pertinent highlights of electron fundamentals. While doing so we shall assume a certain background, and stress only those points which are important to the subject at hand.

The Elemental Charges

Usually, when we think about electric current, we have in mind the electron in motion. While it is true that both the elemental negative charge and the elemental positive charge (the electron and the proton) are capable of motion, the electron, having a mass of about 1/1850 of the proton, is the more mobile of the two and is responsible for *conduction* currents through metals, vacuum tubes, and other radio components. Atoms of gases which bear a preponderance of positive charges because electrons have been removed from the atom, thereby forming *positive ions,* also can move as an electric current, as in the case of the gaseous rectifier tube, but such convection currents are only of limited interest at the moment. It is the electron current which concerns us most.

Although an electron has never been seen, many things about it are known: at least, certain ideas have been advanced, which so far have withstood the onslaughts of critical examination and experiment. Among these is the mass of the electron. In most applications, especially in connection with radio and allied activities, the velocity of the electron is very low. At such speeds, or when at rest, it behaves as if it had a mass of approximately 9×10^{-28} grams. Its mass, however, is not a fixed quantity; instead, it is a variable which depends upon its velocity. If the velocity is less than one-tenth of the speed of light (186,000 miles per second), the change in mass is negligible from that established for it when at rest. From the practical viewpoint, such velocity is obtained with accelerating voltages approximating 2,500 volts. At voltages in excess of this value, the mass increases; at about 10,000 volts, the electron velocity approximates about one-fifth of the speed of light, and at about 1,000,000 volts, which is far beyond anything ever encountered in the conventional cathode-ray tubes, the electron is traveling at about 95 per cent of the speed of light.

Einstein showed that electron velocity can never attain the speed of light because of its increase in mass, which would reach infinity at the speed of light. The change in mass with velocity is expressed by the equation

$$m_a = \frac{m}{\sqrt{1 - \left(\frac{v}{c}\right)^2}} \qquad (2\text{-}1)$$

where m = the mass in grams of the electron at rest
m_a = the apparent mass in grams at the new velocity
v = the velocity in centimeters per second
c = the velocity of light in centimeters per second (3×10^{10}).

If expressed in terms of the accelerating voltage active upon the electron, its apparent mass at the new velocity is

$$m_a = m(1 + 1.94 \times 10^{-6} E) \qquad (2\text{-}2)$$

where m and m_a are as before and E is the accelerating potential expressed in conventional volts[1].

The interest in these figures does not lie in the need for mathematics when learning the operation of the tube, but rather as a means of fortifying the statements relative to the conditions of operation as the electron velocity is increased with high accelerating voltages. Lest we digress too far, let it suffice for the moment to say that the increased brilliance of the trace, accomplished by increasing the accelerating voltage, is related to the apparent increase in the mass of the electron as its velocity is increased. By virtue of the increased speed and mass, the electron is given increased energy, which in turn is transformed into increased brilliance at the point where the beam strikes the atoms of the screen material.

The amounts of electricity represented by the two kinds of elemental *charges* are equal, although far too small individually to be of any practical utility. Specifically, the elemental charge, whether positive or negative, is of the order of 1.59×10^{-19} coulomb. Any useful value of current, no matter how little it may be, usually represents the movement of prodigious numbers of electrons. One microampere of current requires that about 6×10^{12}, or 6 million million, electrons advance past a point in the circuit in one second. This is a fantastic amount, but fortunately one which need not be kept in mind. In the final analysis, we concern ourselves usually with the over-all quantity of electricity which is transported in a unit time, rather than with the number of individual electrons.

Just as a matter of comparison, we might mention that when 6×10^{18} electrons move past a point in one second, one ampere of current is said to flow in the circuit and this corresponds to the transportation of one coulomb of electricity. Thus it is clear to see that the significance of the elemental charge as a certain amount of electricity is negligible. Yet it is important from the viewpoint of study, because the behavior of great quantities of electricity can be studied by noting the behavior of small quantities. In other words, the behavior of electric current can be learned by noting the behavior of a single electron, because all electrons are alike, regardless of origin.

Current Without Conductors

When we think of electric current or electrons in motion, we customarily associate the current with some confining medium or transporting path, as, for example, a wire. This is a commonplace condition in electrical circuitry, yet not necessarily a "must" condition. In other words, it is not necessary that a metallic conducting path be present in order that electrons flow in a directed path and constitute electron current. The operation of a vacuum tube demonstrates that a metallic conducting path is not essential; the grid, plate, and screen current within the tube advance through "space" without benefit of conducting paths. The physical conducting paths appear at the elements and the circuits connected to the elements. The cathode-ray tube is another example of electron flow through space.

These references to conductorless paths for current should not be misconstrued as implying that a so-called open circuit is an operative system. Direct electron flow through space occurs only under proper conditions. For the present, we are concerned solely with electron currents which are identified as conduction currents when they advance through wires, and we take for granted that wire circuits as paths are complete in every respect. We raise this point because the properties of electron currents as associated with wire conducting systems are applicable to like currents which are traveling through space.

Although we speak freely about vacuum-tube currents as being electrons in motion, we seldom think of them as having the properties of current in a wire because we have no occasion to consider these effects inside of many vacuum-tube devices. For example, an important attribute of the electron in motion is a magnetic field which surrounds the moving charge. Seldom do we pay any heed to this phenomenon in a high-vacuum rectifier tube, in an amplifier tube, or in an oscillator tube. Admittedly, the movement of electrons through such devices is not in the form of a concentrated beam; if it were so, there might possibly be a remote association with a similar current through a wire. In the case of the cathode-ray tube, however, the electrons are confined into a narrow path; therefore, there is a closer similarity between a stream of electrons moving through a wire, and one advancing through this tube.

But it is not this similarity which forces us to stress the condition. Instead it is that not only does the electron stream within the cathode-ray tube possess every property of an electron current, especially the associated magnetic field, but also that the operation of the tube depends in part upon several phenomena which are tied in with the magnetic field. Electromagnetic focusing would be impossible without the magnetic attribute of the electron beam. The same, of course, is true as far as electromagnetic deflection is concerned.

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[1] Terman, F. E., "Radio Engineers' Handbook," 1st ed., p. 277, McGraw-Hill Book Co., New York, N. Y., 1943.

The absence of concentration of electrons in a vacuum tube does not remove the electron stream from the conditions described as the basic laws of electricity and magnetism as applied to electric current—those conditions which are, of necessity, associated with the beam of electrons in the cathode-ray tube. Both are governed by identical laws; it is just that these laws are of more immediate interest in the cathode-ray tube than in the ordinary vacuum tube. The electrons advancing through a conventional vacuum tube are surrounded by a magnetic field, but it is unimportant in that case because the ordinary manner of application of such tubes does not require recognition of this field. In the cathode-ray tube, not only does the magnetic field demand recognition, but also it is put to good use.

Stationary Charges

Not all of our dealings with electrons in connection with the operation of the cathode-ray tube are restricted to charges in motion—that is, dynamic electricity. A good deal of what takes place within the cathode-ray tube is the result of the behavior of electrons at rest, of the positive charge at rest. While it is true that we cannot create a positively charged plate without first making electrons move away from that plate, leaving a residual net positive charge, we ordinarily ignore this transient motion of the electron. Conditions are considered only after the electron has come to rest somewhere else, where it sets up a negatively charged surface.

If these statements remind you of "charged" bodies of surfaces, the basis of operation of the conventional vacuum tube wherein the elements act as charged surfaces, or even the principle of operation of a capacitor, your thinking is along correct lines. Many of the performance capabilities of the cathode-ray tube stem from the creation of a "charged" state in some of the components which are part of its structure. The electrostatic fields and lines of force existing between these charged surfaces account for such actions as electrostatic focusing and deflection, because they determine the paths of electrons which pass under their influence.

Taking into account the properties of current and the fields between charged surfaces, we shall see that the basis of focusing and deflection is the establishment of certain paths for the emitted electrons and the beam by these phenomena. To discuss each completely is beyond the scope of this book, yet a review of the basic principles is essential. This can be done most easily by grouping the actions into electrostatics and magnetics and discussing each separately. The discussion will, however, be limited to those subjects within each group which are most directly associated with the title of this text.

Electric Fields and Lines of Force

It is a basic law of physics that unlike charges attract each other and that like charges repel each other. The development of this law consumed time and effort in experiment, observation, and analysis. Originally, when it was noted that bodies bearing charges reacted in a certain manner upon each other, the general approach was that some sort of "action at a distance" was taking place. Among the early investigators, Faraday was a strong objector to this philosophy. Eventually he conceived the idea that something existed "between" the bodies which displayed attraction and repulsion for each other. The result was the concept of electric fields and lines of force as being in the space between the objects.

The graphical presentations of the electric field and the lines of force, are strictly mental devices, to assist in visualizing certain phenomena. Sometimes it is convenient to think of the field lines or the lines of force as having certain physical properties; but these properties are nothing but explanations of what happens, not the cause. By accepting these concepts as a working tool without trying to demonstrate validity, it will be easier to comprehend the subject of fields, and in particular electron lenses.

By an electrostatic field or electric field, the physicist means a special condition of space around every elemental charge of either polarity. By the special condition of space, the physicist means that energy, attributable to the charge, exists around it. Moreover, this energy is capable of exerting a force upon another charge. Thus every elemental charge, whether isolated in space or resting upon the surface of a body, is inseparably associated with an electrostatic field, which may be called a *field of energy*. Or, if we examine the condition from the viewpoint of the force which it may exert upon another charge, then the electrostatic or electric field is a *field of force*.

Since the field is one of energy or force, neither of which is a substance, the field is invisible, and it is detectable only by its effects. Such effects are the forces which the field is capable of exerting upon something susceptible to it, and such a "something" is another charge or charges.

Now, recalling that every elemental charge has a field of its own, and placing a charge A within the field of another charge B, if we say that the field of A is capable of exerting a force upon charge B, we are stating only a part of the truth. In reality, charge B also has a field, so that the field of A is acting upon B, and the field of B is acting upon A.

When we speak about forces being present in a field and being the effect of the field, it must be remembered that this is a mutual condition involving all of the fields due to all of the charges. The manifestation of these forces is the action described by the basic laws of physics—the attraction between unlike charges and the repulsion between like charges.

Now, force not only has magnitude, but is also a *directed* quantity. That is, it acts in a definite direction. (Such a quantity is called a *vector.*) If we say that a field is perceivable by its effects and these effects are forces, then the effects (forces) have both direction and magnitude. Here we have a clue to the means of representing the field as a whole, or rather, a means of illustrating the special condition of either the space around a charge, or the condition of space between charges.

What is done is to use lines which show the direction of action of the forces present in the field. We might describe these lines as being the paths of the force or the paths over which the force would be exerted; if the force is applied to move a charge, the motion of the charge would be along such a line. Hence the physicist's artifice of *lines of force* as the pattern of the electrostatic field. This makes sense and is a very convenient method of presentation because it does inform about the condition of the space.

Admittedly, such a direction line alone is not all the information about the forces present within a field, because it lacks data relative to the magnitude of the force. However, if we are interested only in the general pattern of the field, and not in the exact constants of a specific field, we can do without the magnitudes of force at different points in the field. Such actually is the case; the pattern of the electrostatic field between two unlike and between two like charges is of much more value to us than the actual distribution of forces. This is so because our object now is the presentation of information about the purpose of the lines of force and their behavior, rather than the actual magnitudes. So the stage is set for the picturization of an electrostatic field and lines of force.

The Field Between Unlike Charges

In Fig. 2-4 is shown the electrostatic field between two unlike charges. These need not be two elemental charges of unlike sign; they could just as readily be accumulations of unlike charges—all of one sign in one group distributed around the surface of a sphere, and all of opposite sign similarly distributed on the surface of another sphere. The general nature of the field will be the same for both of these conditions, although it must be said that such a pattern will not hold for all configurations of the body which mounts the charges. When the shape of the field sources is spherical and the charges are unlike, the pattern of the field is as shown in this illustration. A change in shape of a group of charges will cause a different type of field, as will be illustrated later. In the meantime, it is possible to develop the information about the behavior of the lines of force using this elemental field pattern, because the manner in which the lines of force behave is fundamental and not subject to specific configuration either of the field source or the field itself.

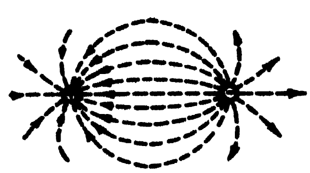


FIG. 2-4. — Electrostatic field between two unlike charges.

Many interesting observations may be made about such a field in terms of the lines of force. For example, some lines of force are shown joining the two unlike charges which we shall assume to be separated by a certain distance. The actual dimension of this distance is unimportant. Other lines are shown incomplete. This, however, is purely a matter of limitations in illustrating, because every line of force between two charges joins these two charges. Therefore, if you can visualize such a thing, the lines of force issue from each charge in all directions and join the charges.

The two charges in Fig. 2-4 are shown relatively close to each other. Such, however, need not be the case; the unlike charges can be far apart, without in any way changing the condition that lines of force from a negative charge will terminate on a positive charge, or vice versa. Theoretically, the amount of separation of the unlike charges is of no consequence.

Then there is the matter of the space between the lines of force. This, too, is a limitation in illustrating. Actually there is no space between them; all the area, in all directions, is filled with lines of force. In drawing the field, we select certain lines only as being representative, and show them.

As to the direction of the field, as shown by the arrowheads, it is the direction in which the force would be exerted *along the line of force* in accordance with a certain convention. This requires some elaboration. Two conventions exist for the direction of an electrostatic field. One frequently used convention assumes such direction of the field as would be indicated if a test charge of positive polarity were placed into the field. The other convention, which is used in this book, assumes such direction of the field as would be indicated if the test charge placed into the field were of negative sign. Our preference for this convention is based on the fact that the electrostatic fields in the cathode-ray tube act on the electron, which has a negative charge; therefore, we avoid confusion by considering that direction which conforms most closely to the final analysis of cathode-ray operation.

Accordingly, the direction arrows point away from the negative charge and toward the positive charge, and show the direction in which an electron would move when under the influence of the field. In this connection, we might mention that the field shown in Fig. 2-4 does not require the negative charge to advance to the positive charge. It is possible to imagine these two charges as being fixed in their respective positions, in which case conditions would remain constant and an electrostatic field of a certain direction, which *could* cause a certain motion on the part of an electron placed within the field, would be effective.

Behavior of the Lines of Force

Continuing with our discussion of the field between two unlike charges, as shown in Fig. 2-4, several very important comments must yet be made. Perhaps, from the viewpoint of utility, these are more important than any made so far. We mentioned earlier that the effect of the field was the application of forces to those things which were susceptible to such forces. Let us consider for the moment the basic law which states that unlike charges are attracted to each other. Such attraction often involves physical motion and, according to our convention, it would be a motion of the electron.

Is such motion a function of the charges themselves, that is, their presence? Or, put differently, do the charges move of their own volition because they are of unlike polarity? According to theory, the answer is negative. It is true that the polarity of the charges determines the field, but the effect of the field as interpreted in the motion of the charges is due to the behavior of the lines of force. As stated earlier, these lines are aids to visualizing what actually does happen, and have no independent existence outside of the imagination. Since they are imaginary, we may assign any properties we choose to help describe what actually happens. One such property is that of lengthwise contraction; we assume that lines of force tend to shrink along their length. For example, a line extending from a positive charge to an electron, in shrinking, would draw the electron toward the charge. If the electrons were carried further away from the positive charge, the line of force would be stretched; a mechanical force would be required to do this. Given the opportunity, as determined by the mobility of the charges, the electric lines of force between the two charges will bring the charges together so as to satisfy the fundamental condition of minimum length of the line of force.

It is because of this, as well as a few other conditions, that a comparison is made between the behavior of such lines of force and that of stretched rubber bands. Of course there are limits to the application of this analogy, but it is a reasonable one when we visualize the line of force as possessing energy which can be put to use, just as the stretched rubber band possesses energy which can be put to work. Most certainly the tendency of the stretched rubber band is to shorten itself, and in so doing, to do work. The same can be said about the electric line of force. Given the opportunity to contract, the energy present along the line of force will be transformed into other forms of energy—perhaps kinetic energy applied to the electron so as to get it to move, and of course, into heat. At any rate, the energy in the field, distributed along the lines of force, is capable of doing work during the time that the lines of force contract.

Another property of the lines of force is that lines active in the same direction appear to repel one another laterally. This accounts for the curvature of the lines of force which join two charges at points other than on the straight-line axis between the charges. The line of force along the axis is straight because the amount of repulsion, from other lines laterally positioned, is the same on all sides; in other words, the forces are symmetrical. However, at distances from the axis, the lines of force curve outward because of the greater amount of repulsive force originating from the direction of the axis and less repulsive force originating from the outer side of the force line. Nevertheless, spaces still do not exist between the lines of force; the lines simply curve outward instead of being straight.

This property of the lines of force between two unlike charges may not have much bearing upon the behavior of such charges. Being subject to attractive forces, it is the shortening of the line of force along its length which is the paramount action. However, the lateral repulsion between the lines of force with like direction is very important in the behavior of like charges, which are dealt with in the paragraphs which follow.

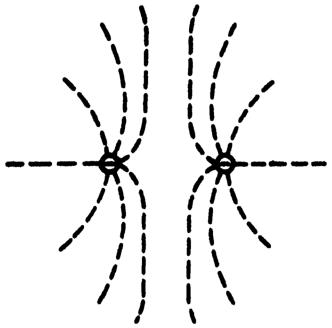


FIG. 2-5. — Electrostatic field between two like charges.

The Field Between Like Charges

Fig. 2-5 shows a pattern of the electrostatic field between two like charges. Whether they are two positive or two negative charges is immaterial; we show two of the latter simply as a matter of choice. If they were two positive charges, the field would still be the same.

Let us examine this field pattern somewhat critically. The first condition we note is that there seem to be no termination for the lines of force. Why not? The answer is simply that we show only a portion of the fields associated with these two like charges. In reality, each of these two like charges has its own field with a charge of opposite sign located at some remote point. It is to these unlike charges that the lines of force shown in the figure continue. In other words, when we show the field between two like charges, we really are showing only that section of each of two fields, extending between two unlike charges, which is in the proximity of the two like charges.

It is conceivable, in theory, to visualize these two like charges isolated in space, each with its field extending to infinity, and to describe a direction for each line of force according to its effect upon a "test" electron (one which is placed in the field for determination of lines of force). In accordance with this idea, the directions of the lines of force which face each other would be the same. However, the reason for tying in the two like charges under consideration with two other charges of opposite sign is that it permits the closest correlation with what is to follow during the discussion of the cathode-ray tube. Fig. 2-6 is the drawing we shall use for discussion from this point on; this is the same, in substance, as Fig. 2-5, with the addition of the "remote" charges.

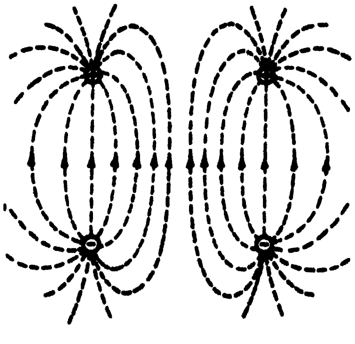


FIG. 2-6. — Combined electrostatic field between like and unlike charges.

A few differences exist between the two drawings, but in the main these are due to perspective, rather than to any actual difference in conditions. What we have done is to reorient ourselves relative to the imaginary lines of force. The fact that we illustrate two pairs of like charges instead of a single pair does not complicate the discussion. Whether it is a single pair of like charges or two pairs, the question remains the same—"What is the basis of the repulsion between like charges?" Remembering that the attraction between unlike charges is attributable to the contracting properties of the lines of force, it is not too far-fetched to suppose that the repulsion likewise is due to some action on the part of the lines of force.

This is the case; it is the lateral repulsion between lines of force having like direction, which is responsible for the repulsion between charges. As is evident in the illustration, the direction of the lines of force facing each other and which join the two pairs of unlike charges is the same; these lines feel repulsive forces between them. Being attached to the lines, the charges likewise feel the same forces and move apart. It is necessary to understand that the repulsion between similarly directed lines of force does not depend upon motion of the charges. In fact it is the reverse; the motion of the charges is due to the repulsive forces. These exist even if motion of the charges responsible for the fixed field cannot take place because of the nature of the system.

Relative to the pattern of the field shown in Fig. 2-6, another significant condition must be mentioned. This is the flattening of the facing lines of force as the consequence of the repulsion between them. While this may not have a great meaning with respect to the actual action, the change in the configuration of the lines of force so that they run almost parallel to each other is a very important point. To state why this is so at this time would be premature in the organization of this discussion. All that we shall say here is that we should bear that condition in mind for future reference.

At this stage you may be wondering about the connection between the field as shown in Fig. 2-6 and what happens in the cathode-ray tube. The justification for showing the relationship between like charges is twofold. First, the type of configuration of the facing lines of force shown in Fig. 2-6 is a very close approach to what will be found in the cathode-ray-tube electrostatic focusing system. The second reason is to explain the background of the action whereby an electron moves away from a negatively charged body. This could not be done very well using the field between two unlike charges.

On the whole, the field patterns shown in Figs. 2-4 and 2-6 serve well as background for the field patterns which will be found not only in the electrostatic focusing systems, but also in the deflection systems. Much more could be said, but it will be held in abeyance until later, when the expansion of these basic ideas will permit a much closer correlation with actual conditions of tube performance.

Charged Bodies

Expanding upon the basic conditions which we have described, it stands to reason that if a body or a surface is made to bear a preponderance of a charge of a single sign, that body will, in principle, display the same electrical effects as a single charge of the same sign. Of course, the magnitude of the effect will naturally be much greater, because there are many more charges present. Therefore, if we arrange one surface to bear a preponderance of electrons and another to bear a preponderance of positive charges, and these two surfaces are placed so as to face each other, but are separated, an electrostatic field will exist between the two charged surfaces. The space between the surfaces will be filled with lines of force which will have a certain direction.

Such a system is shown in Fig. 2-7. The voltage source *B*, with an arbitrary difference of potential between its terminals, is connected to two parallel metal plates *A* and *C*. Prior to the application of the voltage, a condition of electrical equilibrium existed on the two plates; equal numbers of unlike charges resided on each of the plates, therefore, the space between the plates was devoid of any field.

When the voltage is applied, this two-plate "condenser," with air separation between, becomes "charged" like any ordinary electrical capacitor would.

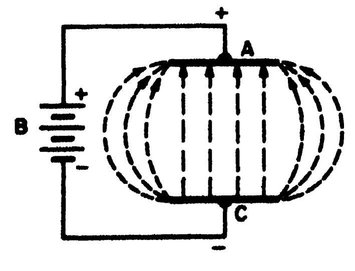


FIG. 2-7.—Electrostatic field between two charged plates.

One of the plates *A* bears a preponderance of positive charges and the other plate *C* bears a preponderance of negative charges by virtue of the polarity of the charging source; an electrostatic field is created between the plates and the space is filled with lines of force. In view of what was said earlier, we need not comment on the origin of the lines of force; if anything deserves comment, it is the configuration of this field as compared to the field between two unlike charges as previously illustrated.

Between the plates, the field is made up of lines of force which are straight, and we may assume that they are uniformly distributed within the boundaries of the plates because the separation between the plates is small compared to the other dimensions of the surfaces. If we neglect the conditions near the borders of the two plates, the number of lines of force penetrating any unit area of surface on the plates will be the same everywhere on the plate. Such a distribution of lines of force constitute a *uniform field;* this condition would not hold if the plates were not parallel.

As to why the lines of force which join the charges on the two plates are straight, rather than both straight and curved as shown in Fig. 2-4—that too is simple to explain. To begin with, charges made to reside upon a flat surface will normally distribute themselves uniformly over the surface. With large numbers of charges residing on each of the plates, there are many lines of force. Each of these, with the exception of the fringe at the boundaries of the plates, feels equal to repulsive forces all around it. With forces of equal magnitude tending to repel each line from all directions, the balancing of these forces leaves the line straight.

The curvature of the lines of force at the borders already has been explained in discussing the two charges of Fig. 2-4. These lines curve outward because the repelling force issuing from the area between the two plates exceeds the force which stems from the area on the outside of the lines. The result is outward curving lines of force. This nonuniform fringe field usually can be neglected.

Regarding the direction of the field, we follow the previous convention, namely that direction as would be defined by its action upon an electron. Therefore, the line of force points away from the negatively charged plate and toward the positively charged plate.

Energy of the Field

Several very interesting and important observations can be made about such an electrostatic field. First, the creation of it requires work, or energy, because the separation of negative charges from positive charges is accomplished only by the application of some force.

Moreover, as more and more electrons are removed, the attractive force of the remaining charges upon the electrons becomes greater and greater. In like manner, the accumulation of electrons on the other plate also becomes increasingly difficult, since like charges repel; therefore, force must be exerted to push these negative charges onto the so-called negative plate. As more and more electrons are piled onto this plate, the repulsive force naturally increases and must be overcome in order to push more electrons onto that surface. In short, *work must be done upon these electrons in order to get them to redistribute themselves in this manner.*

Bearing in mind that these electrons were initially at rest, some source must impart energy to them so as to cause them to move around through the circuit in a specific manner and specific direction—against the natural forces which either bind them to each other or tend to make them repel each other, depending upon the polarity of the charges. This function of supplying energy to the electrons is performed by the device which "charges" the system. In Fig. 2-7, this is the battery *B* which can be replaced by a voltage source of some other type, such as a vacuum tube or a generator, without altering the basic phenomena. The voltage source introduces the initial *difference of potential* or *electromotive force* (emf), which exists between its terminals, necessary to make the electrons move around the circuit.

As the negative charges are transferred from one plate to the other, they convey energy and gradually establish a corresponding difference of potential between the plates themselves. The energy which the voltage supply imparted to the electron is transferred to the electrostatic field (between the plates) when the electron comes to rest. Therefore, we can view the difference of potential between the two parallel facing surfaces as representing the capabilities of the electrostatic field to make electrons move, just as the initial difference of potential between the terminals of the voltage source enabled electrons to move through the system initially.

The electrostatic field between the plates can make electrons move either between the plates, or through a conductor which joins the two plates. The limit of movement of the negative charges under the influence of the original voltage source is reached when the difference of potential established between the plates equals the difference of potential between the terminals of the battery. At this time, the magnitude of the attractive force of the positive charges on the positive plate and the magnitude of repulsive force of the negative charges on the negative plate are sufficient to offset the electromotive force (emf) represented by the difference of potential, or voltage, of the battery.

The time required for the difference of potential between plates *A* and *C* to rise to the value of the voltage source *B* is a variable depending upon several factors. None of these are of interest in connection with the systems used in cathode-ray tubes, whose operation is now being explained. Practically, we can assume an instantaneous rise in potential, between the plates, to that of the voltage source *B*. Of course there always will be some time lag, but this is so slight from the usual practical viewpoint that it can be neglected over any range of frequencies presently encountered in cathode-ray-tube applications.

The development of a detectable electrostatic field between the plates, or the setting up of lines of force between the plates, commences with the first electron which is transported, and the field becomes increasingly stronger (greater number of lines of force) as more and more electrons are removed from the positive plate and added to the negative plate.

The ability of the electrons, accumulated on the negative plate, to hurdle the gap separating the plates is strictly a function of the intensity of the field; this, in turn, is a function of the amount of energy which is given the electrons by the so-called "charging" voltage source. If this voltage were gradually increased from zero, a value would be reached when so many electrons had been redistributed that the mutual attraction between the positive charges on one plate and the electrons on the other plate would literally tear away the electrons from the negative surface, in a direction along the lines of force, and a "current" would flash across the gap. This current, or spark, would of course be momentary, but would nevertheless dissipate the entire field.

A phenomenon of this type is an example of *work being done on the electron,* the action being across the space between the plates, rather than through any metallic conducting path around the plates. Since it is possible for the field to exert sufficient force upon the electrons on the negative plate so as to pull them to the positive plate, it stands to reason that if electrons were positioned *between* the plates they could be acted upon in similar manner. The motion of these space electrons would be governed by the forces present in the field, which in turn would correspond to the field intensity at the point where the electrons were located. This leads to the description of the field from the viewpoint of the forces present at various points in the field.

Forces Acting On Spatial Electrons

Let us use, as the basis of our discussion, the illustration in Fig. 2-8. This shows a uniform field existing between two parallel plates. We shall locate three electrons, *a*, *b*, and *c*, at three different points in the field. One of them, *a*, will be near the negatively charged plate; electron *b* will be midway between the two charged plates; and electron *c* will be near the positively charged plate. All of them will feel the same pull toward the positive plate because of the direction of the field, and, what is more, all of them will experience the same force making them move to the positive plate. Naturally, all will advance along lines of force.

The movement of the electrons is not due solely to the attraction by the positively charged plate, *A*. A distinct contribution to this action is made by the negatively charged plate as well. During the time that the positively charged plate is *attracting* the electron, the negatively charged plate is *repelling* the electron toward the positive plate. Hence two forces, in the same direction, are acting upon the charge.

If we imagine the electron located midway between the plates, in this case electron *b*, equal forces of attraction and repulsion will move the charge in a single direction, i.e. toward plate *A*. The total force moving the electron will be the sum of these two individual forces, both, you will remember, acting simultaneously in the same direction.

If the location of the electron is changed so that it is no longer midway between the plates but near the negatively charged plate, like electron *a*, the total force acting upon the electron will still be the same as before. Although it is true that the magnitude of attracting force has been decreased due to the greater separation between the charge and the positive plate, the magnitude of the repelling force has been correspondingly increased due to the greater proximity of the electron to the negative plate, thus maintaining the total force constant.

Reversing the location of the electron, that is, locating it near the positively charged plate *A*, as in the case of electron *c*, does not change the force acting on the electron. The reduction of repelling force due to the increased separation between the charge and the negatively charged plate is compensated for by the corresponding increase in attractive force from the closer positively charged plate.

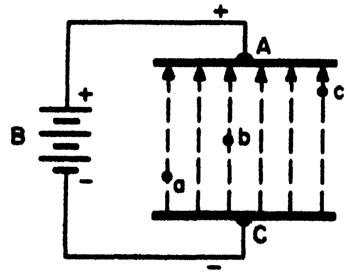


FIG. 2-8.—Electrons placed in electrostatic field between two charged plates and acted upon by force of the field.

Thus it is understandable, even without a mathematical analysis, that the electrostatic field acting upon the electron anywhere within such a uniform field is constant. This holds true for any one set of conditions which establishes the magnitude of the field and for any one set of plates separated by a fixed distance.

The actual force in dynes experienced by the electron is of no consequence in this discussion. It can be small or it can be great, depending upon the difference of potential between the plates *A* and *C*. The important detail to bear in mind is the direction in which the force acts, and that the force is constant throughout the field.

Velocity of the Electron

In connection with the transverse motion of the electron through the space between two plates, across which a difference of potential exists, it might be well to consider the velocity for future reference. In Fig. 2-8 we assume several electrons at rest, placed within the field. Because of the action of the field upon these electrons, they will move toward plate *A*. In other words, the *field is doing work on* the electrons and accelerating them. This is equivalent to saying that the electron acquires energy—kinetic energy, the energy of motion. To specify the magnitude of energy thus gained by the electron, it is customary to express it in terms of *electron volts of energy* or simply *electron volts*. This conforms to a previous reference that voltage was indicative of the energy possessed by an electron. Consequently, if an electron which is located between two plates across which a difference of potential of 300 volts exists, and the field accelerates the electron from the point of zero potential to the point of 300 volts potential, or through 300 volts, the electron has gained 300 electron volts of kinetic energy. This is also described by saying that the electron has fallen through 300 volts, analogous to a falling object which loses potential energy and gains kinetic energy during the fall. Thus, if an electron falls through 60 volts, it has gained 60 electron volts of energy, or if the energy possessed by a moving electron is described as being 3,000 electron volts, it means that the electron has been accelerated by, or has fallen through, 3,000 volts.

Quite frequently, the velocity of an electron is expressed in terms of electron volts, or more simply, volts. Thus if an electron were described as having a velocity of 300 volts, it would mean the velocity which it acquired when falling through 300 volts, or when accelerated through 300 volts. The approximate velocity corresponding to a number of volts is expressed by the equation

$$v = 5.93 \times 10^7 \times \sqrt{E} \qquad (2\text{-}3)$$

where v is expressed in centimeters per second, and E is the difference of potential responsible for the acceleration of the electron.

This equation is applicable to accelerating voltages of several thousand volts, at least to values which will not cause velocities which exceed about fifteen-hundredths that of light, where the change in mass of the electron due to its velocity starts to become significant. The final velocity of the electron when advancing through a field along the lines of force is of interest for several reasons. First, the equation shows that the final velocity is independent of the spacing between the plates and is proportional to the square root of the voltage across the plates. Second, the action of electrostatic deflection involves the matter of electron speed, deflection voltages, and frequency of operation.

Equipotential Points and Lines

According to convention, when an electron advances from point *A* to point *B* through a circuit, as in Fig. 2-9, the charge has moved from a point of lower potential to a point of higher potential. To make the charge move through the circuit requires that work be done on it, which means that energy must be imparted to it. The source of this energy is the voltage supply, and the movement of the charge is accounted for by the fact that the voltage source sets up a difference of potential at different points throughout the system, in accordance with the constants of the system.

At any point in the system, the potential of that point is a measure of the work which must be done upon the electron to make it advance from another point to the point in question. Therefore, it is possible to characterize a point in a system such as *B* in Fig. 2-9, by reference to its potential with respect to some other point—as, for example, a zero-potential reference point, such as *A* in the same illustration. When this is done, the direction of advance of the charge (current) in the circuit is indicated; also, the electrical level of that point in the system is stated.

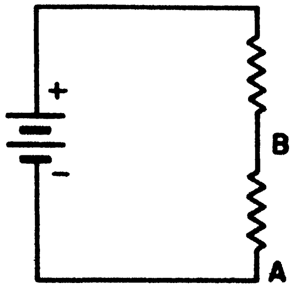


FIG. 2-9.—In circuit shown, electrons move from *A* to *B* because of electromotive force (emf) supplied by battery.

The use of the term "potential" as an indicator of the electrical level, as a measure of the work which must be done upon a charge to get it to that point, and finally as a means of indicating the direction of movement of the charge, is not restricted to electrical circuits; it can just as readily be applied to points in space. As a matter of fact, such reference to the potential of points in space is very essential in nonuniform structures, because it affords a means of indicating and describing the electron paths more conveniently than when the lines of force are shown.

By definition, "the potential of a point is the work which must be done on a unit charge in order to transport it from a ground reference point to the point in question". Since work is the product of force by distance, the potential of a point in space is proportional to either force or distance when the other quantity is constant. We stated before that the force anywhere in a uniform field is constant at all points, which means that the potential of any point in a uniform field is proportional to the distance from the reference point.

For example, in Fig. 2-10 we show the uniform field which exists between two surfaces one centimeter apart, with a difference of potential of 300 volts between the two plates. If we select the negative plate as our reference point, the potential along any line of force between the two plates will increase progressively in whatever increments we may select from zero at the reference point (the negative plate) to the maximum (300 volts in this case) at the positive plate. The work which must be done on a charge in order to move it

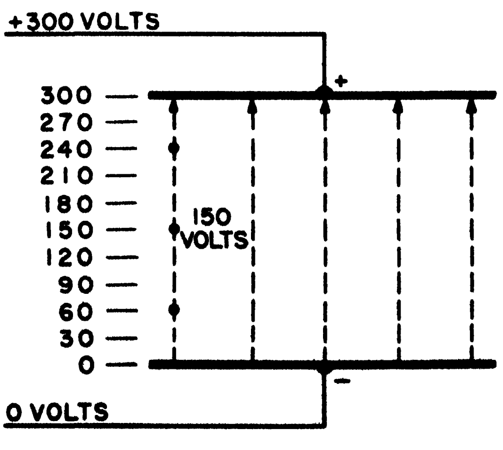


FIG. 2-10.—Uniform field which exists between two charged plates (fringing disregarded). Point mid-way between plates will have potential of 150 volts, as shown.

from the reference point to another point midway between the two plates along a line of force, amounts to half of what would have to be done to transport it to the positive plate. Consequently, a point half-way up along the lines of force would have a potential of 150 volts.

If we divide the distance between the two surfaces into equal intervals of 0.1 cm along a line of force, we would have point of increasing potential 0.1 cm apart. Each 0.1 cm would correspond to an increase of 30 volts, since the total distance of 1.0 cm, equal to a difference of potential of 300 volts, is divided into ten equal parts. The first 0.1 cm point upward from the reference point of zero potential corresponds to 30 volts; the 0.2 cm point corresponds to 60 volts, and so on. If the maximum difference of potential between the plates were 3,000 volts, and the separation between them were 1.0 cm, then each of these 0.1 cm points would correspond to an increment of 300 volts. For any difference of potential, an infinite number of incremental points can be established.

We said that in a uniform field, all the lines of force are alike, neglecting the fringe field, and therefore, points of different potential as established on one line of force are the same on the other lines of force. In other words, points of equal potential exist on each line of force, and if we join these equipotential points, we develop a pattern of *equipotential layers or surfaces,* each of which may be identified numerically in practical volts. In this example, shown in Fig. 2-11, the equipotential surfaces increase in steps of 30 volts from the zero-voltage reference point. A clearer representation of equipotential mapping of the field appears in Fig. 2-12, from which the lines of force have been omitted. From the viewpoint of information, the illustration in Fig. 2-12 furnishes just as much as Fig. 2-11. As stated before, the number of equipotential lines which can be set up is infinite, being determined only by the increment selected.

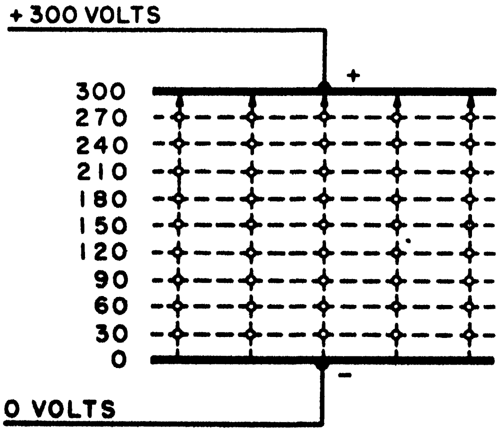


Fig. 2-11.—Points of equal potential along lines of force between charged plates are connected to form equipotential lines.

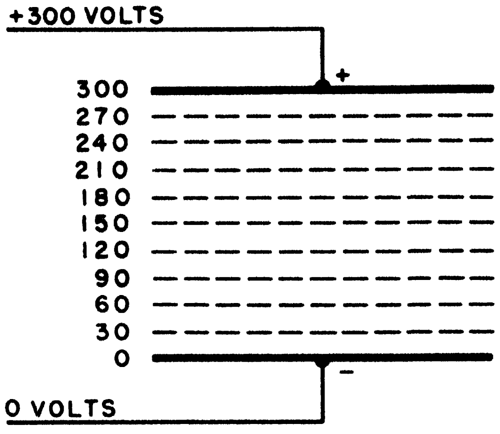


Fig. 2-12. — Equipotential surfaces between two charged plates, neglecting the fringe field.

Let us analyze Figs. 2-11 and 2-12. Referring first to Fig. 2-11, take note of a very significant relationship—the equipotential surfaces *cross the lines of force at right angles.* Since the line of force is the path along which the electron would tend to travel, unless some other simultaneous force also is active (which is not the case in this example, because we assume the test electrons starting from rest), it is possible to make the general statement that a *charge tends to move across equipotential surfaces at right angles.* The actual path is, of course, affected by the tendency of a particle to keep moving in a straight line, if it enters the field at high velocity. This is a very important condition to bear in mind, because in very many instances, especially in electron lenses, the lines of force existing between the charged surfaces do not form uniform fields. Consequently the equipotential surfaces are not parallel, as they are in this simple configuration. Instead, they may be curved to varying degrees, and it is essential that you understand this right-angle relationship in order to appreciate why the path of the electrons is such as to make them diverge or converge, depending upon the type of curvature of the equipotential surface.

Because of this basic condition, the simple equipotential pattern of Fig. 2-12 furnishes much information. Obviously, the numerical identification of the potential levels is pertinent to the recognition of the direction of travel of the electron, although if only the polarity of the plates were given, it would serve the same purpose. This, however, is true only in the simple case of uniform fields. When the field is not uniform, the

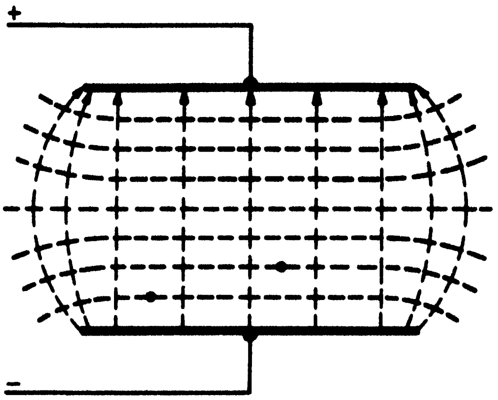


FIG. 2-13.—Equipotential surfaces and electrostatic lines of force between two charged plates, taking fringe field into account. Electrons positioned between plates will move in the direction of the arrows.

polarity designation is insufficient, as is shown in Fig. 2-13, wherein the pattern of Fig. 2-12 has been modified by the addition of the equipotential surfaces *within the fringe portion of the electrostatic field* as shown in Fig. 2-5. (Observe that because the equipotential surfaces are seen edgewise in the figure, they must necessarily be drawn as lines.) Compare Figs. 2-12 and 2-13, especially the separation between the equipotential surfaces in the uniform portion of the field, and in the nonuniform field at the boundaries of the plates. The difference is brought about by the fact that the equipotential surface *must cross the lines of force at right angles,* and curvatures of the lines of force in the fringe field have varying radii.

Suppose, for example, that we imagine several electrons positioned at rest within the uniform portion of the field and indicate these in Fig. 2-13 by the small circles. How would these charges move? We know that electrons normally advance from the point of low potential to the point of higher potential. This immediately indicates the general direction of electron motion. Next, we observe that the equipotential lines between the plates are parallel to each other, and since the electrons would cross these lines at right angles, we know by simple geometry that the paths of the electrons would be parallel to each other as they advanced toward the upper plate, as indicated by the arrows. This reasoning neglects any deviation in path due to mutual repulsion between the electrons.

Examining Fig. 2-13 further, one might ask about possible motion of the electrons *along* an equipotential surface. Unless propelled that way, the electron would feel no force tending to pull it in that direction for the very simple reason that *all points on an equipotential surface are at the same potential.* If accelerated along a path which is parallel to the equipotential line, that is, perpendicular to the lines of force, the electron will advance in that direction, but it will still feel a force (the lines of force in the field) which will tend to make it cross the equipotential lines at right angles. How successful the electron may be in accomplishing the latter depends entirely upon the relation of the forces, that is, the accelerating force parallel to the equipotential surface and the force at right angles to it, along a line of force. The path that the electron would describe between the plates would result from the joint effect of these two forces. This will be discussed again later.

Although the equipotential surfaces in Fig. 2-13 are in the main parallel, it does not mean that such conditions prevail at all times regardless of everything else. In this case it is strictly a matter of a uniform field existing between the two parallel plates and a small fringe field. If the field is nonuniform, as for example in the case of differently shaped surfaces, shown in Fig. 2-14, the lines of force are not parallel; consequently, the equipotential surfaces cannot be parallel to each other. In the final analysis, the relative configuration of the equipotential surfaces is a function of the relative configuration of the field.

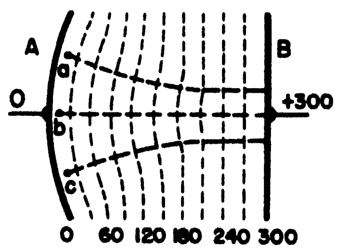


FIG. 2-14. — Equipotential surfaces between dissimilar plates. Electrons tend to cross equipotential surfaces perpendicularly and converge on plate *B*.

Referring to Fig. 2-14, one boundary surface or conducting plate is spherical, whereas the other is plane. These two are charged to a potential difference of 300 volts and the pattern of the equipotential surfaces is as shown. In the vicinity of the curved surface, the lines are likewise curved, gradually straightening out as the plane surface is approached. In the vicinity of the plane surface, the equipotential surfaces run parallel to each other and to the surface. How will electrons behave in such a nonuniform field?

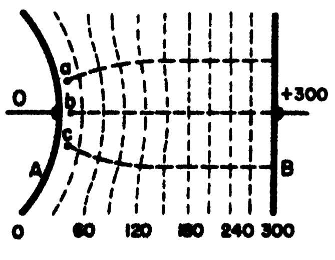
Suppose we locate three electrons, *a*, *b*, and *c*, near the curved surface. These electrons can be considered at rest. Because of the direction of the field, determined by the polarity of the charged surfaces, we know that the general direction of advance will be from plate *A* toward plate *B*. Recalling the basic law that the charge will cross the equipotential line at a right angle, electron *a* will advance toward plate *B* along a *curved* path, determined by the curvature of the equipotential lines, and then will straighten out as it crosses the parallel equipotential lines.

Electron *b*, being located at the center of the spherical surface, will, by the symmetry of the configuration, traverse the space in a straight line, which, of course, crosses each surface of equipotential at right angles. Electron *c*, will, like electron *a*, follow a curved path as it crosses the equipotential surfaces near the spherical surface, because these lines are curved and the right-angle relationship must be maintained; it will then straighten out as it crosses the parallel flat surfaces of equipotential. Thus, three electrons located at three different positions at the spherical surface have been made to converge upon a plate which attracted them. It is readily understandable that if, instead of three electrons, we had many electrons positioned along the curved surface, all would progress to plate *B* in the manner described, and would strike at points which would be very close together. In a way, this is a *focusing* action, although more remains to be said.

Referring once more to Fig. 2-14, it is interesting to note the relationship between the contour of the equipotential lines and the resultant path of the electron. We have seen how a set of straight or plane equipotential surfaces results in a straight-line advance of the electrons. Such movement of the electrons also can be described as being along parallel paths. In the illustration in question, the equipotential surfaces, when viewed in the direction of advance of the electrons, are convex near the curved plate *A*, then gradually straighten out and become plane as plate *B* is approached. In the region where the convex equipotential surfaces exist, the electron motion at right angles to the equipotential lines causes them to converge and then follow a straight and parallel path as they cross the plane equipotential surfaces.

Under the circumstances, it should not be difficult to visualize a change in the shape of the equipotential surfaces so as to make the electrons *diverge* instead of converge. Such a configuration is illustrated in Fig. 2-15. Here are two charged plates, one curved, and the other a plane surface. When viewed in the direction of advance of the electrons through the field, the curved equipotential lines are concave, then gradually straighten out and become parallel. Following the basic law of electron motion across equipotential lines, the three electrons, *a*, *b*, and *c*, will diverge at the outset and then follow parallel paths, with the final result that when they arrive at plate *B*, they will be further apart than when they left the vicinity of surface *A*.

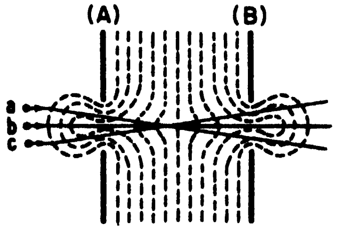
Fig. 2-15. — Equipotential surfaces between dissimilar plates. Electrons, tending to cross equipotential surfaces perpendicularly, diverge.



In the commercial electron guns, which will be described later, the shape of the charge surfaces differs greatly from those which we have shown. This is unimportant to us, as long as the shape of the equipotential lines approaches the fundamental patterns we have illustrated. Regardless of how the equipotential surfaces are obtained, the stated fundamental law of electron advance through the surfaces will hold. There may occur slight modifications as a result of the existence of other forces, but it will not in any way invalidate what we have said concerning the basis of operation.

It might be well to illustrate one more basic condition before we examine the formation of the electron lens. It may prove helpful in more completely correlating the basic conditions with the result. In our explanation of the electrostatic field between the two charged plates, we illustrated a uniform field with parallel equipotential lines between them. (See Fig. 2-12.) Let us now make an aperture in each of these plates at the center, as shown in Fig. 2-16, wherein each plate is a disc with a hole in the center.

Fig. 2-16. — **Equipotential surfaces between two disks, each having an aperture at its center. Basic electron lens is formed.**



For the sake of simplicity in drawing, we show the pattern of equipotential surfaces rather than the lines of force. The interesting detail is that these equipotential surfaces bulge through the holes causing some field to exist outside of its normal region. Therefore, there will be an effect upon the movement of the electron in limited areas *outside* of the plates, as well as in the space between them. If we imagine three electrons, *a*, *b*, and *c*, positioned to the left of the aperture in plate *A*, electrons *a* and *c* will be made to converge as they cross the convex equipotential lines which bulge through the opening in plate *A*, and will tend to *cross* *each other* within the field between the two apertures, and then continue along a diverging path as they cross the concave equipotential lines which bulge through the opening in plate *B*. Electron *b* will continue in a straight path, because it is on the axis of symmetry and advances across each of the equipotential lines at right angles.

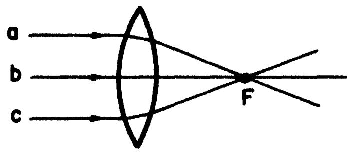


Fig. 2-17. — Simple optical lens.

In a way, the assembly of Fig. 2-16 constitutes an electron lens system. It is similar in elementary fashion to the optical system shown in Fig. 2-17, wherein the three incident light rays, *a*, *b*, and *c* have been brought to a single point at *F*, which corresponds to the crossover point in the space between plates *A* and *B* in Fig. 2-16. While it is true that any set of curved equipotential lines, like those in Figs. 2-14, 2-15, and 2-16, form the equivalent of a simple electron lens, because they refract (bend) the electron path — just as an optical lens refracts a light ray — the complete electron gun system is somewhat more elaborate. Based upon what has been said and what has been shown, it should not be too difficult to imagine the possibility of a combination of charged bodies, such that any desired configuration of equipotential surfaces can be formed; in that way, the path of the electron can be determined and controlled.

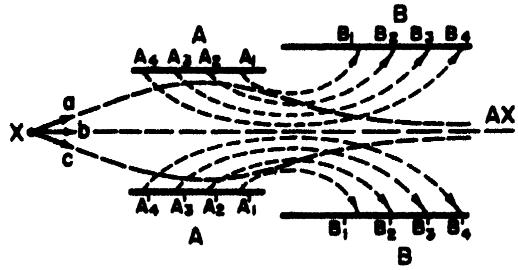


Fig. 2-18.—Electrostatic field between two charged cylinders.

With this in mind, let us examine Fig. 2-18. Here we have two metal cylinders symmetrically arranged along a single axis, or coaxially symmetrical. Both cylinders are made positive with respect to a common reference point, but cylinder *B* is at a higher positive potential than cylinder *A*, with the result that an electrostatic field exists between the two cylinders. A map or representation of the *lines of force* between the inside surfaces of these two cylinders is shown. The pairs of letters indicate the start and end of each line of force and the arrow heads indicate the direction of action of each line of force as being from the surface of lower potential to the surface of higher potential, the direction in which the line would exert force upon an electron.

When we examine the illustration, certain conditions associated with the lines of force pattern come to light. The presentation does not permit a complete picture of all the lines of force which fill the entire space within the two cylinders. Moreover, all the lines of force are active in the same direction; consequently, they tend to repel each other. This condition results in *flattening* of the lines of force nearest the axis. At this point they tend to run along almost parallel paths. In general, the shape of the lines of force is such as to result in equipotential surfaces which assume three general forms — convex within cylinder *A*, substantially plane in the zone where the edges of the two cylinders are near each other, and then concave in cylinder *B*, as shown in Fig. 2-19.

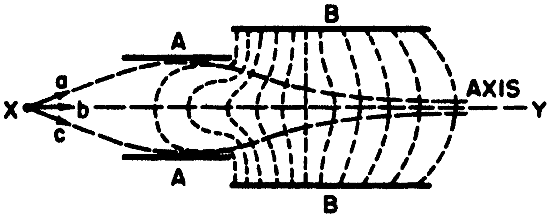


Fig. 2-19.—Equipotential surfaces resulting from electrostatic field shown in Fig. 2-18.

As to the potential distribution within the field, this too is illustrated in connection with each equipotential line. Since cylinder *A* is at a fixed potential with respect to the reference point, the lowest equipotential line shown is that slightly higher than the potential of the cylinder with respect to the reference point. In like manner, the highest potential shown in cylinder *B* is that which approximates the potential of that cylinder with respect to the reference point. It would be possible, if so desired, to illustrate the potential of each equipotential line in terms of the difference of potential *between the two cylinders*, in which case cylinder *A* would be considered as being at zero potential, and cylinder *B* at a positive potential with respect to *A*. The method used is, however, preferable, since it will conform more closely with conditions existing in practice, namely that the potential of cylinder *A*, with respect to the reference point, is also active in accelerating the electrons through the opening of the cylinder.

Point X in Fig. 2-19 is a source of electrons, from which three electrons have been emitted at comparatively slow velocity. One of these, b, advances along the axis X-Y, whereas the other two, a and c, also move forward, but at an angle with respect to the axis. As the three electrons, a, b, and c, advance from point X, they will feel two forces. One of these is the attractive force due to positive voltage applied to the two cylinders, which will urge them toward the cylinders; the second force will be that of the lines of force between the two cylinders, which is away from the inner surface of cylinder A, and toward the inner surface of cylinder B (Fig. 2-18). However, the electrons are not in the field of cylinder B long enough to be attracted to its inner surface, due to the velocity acquired under the influence of the voltage on cylinder A as well as that on B. Moreover, since the normal path of the electrons is through the equipotential surfaces at right angles, and the point of highest potential is farthest removed from its starting point, the electron trajectories are diverging as they cross the convex equipotential surfaces within cylinder A, and change to converging paths as they enter the zone of concave equipotential lines. This is shown by the dotted lines. Once within this zone, the degree of convergence gradually decreases, but since the curvature of the field has already established the main path, the electrons emerge from the field of cylinder B, along approaching paths, and come to a point at some predetermined distance from cylinder B. Although three electrons is an extremely small quantity compared to the actual numbers of electrons in a practical beam, the focusing action and beam formation, when great numbers of electrons enter the focusing field, remain the same.

Practical Electrostatic Focusing System

If we put together the results derived from the discussion so far, we can obtain a typical electrostatic focusing system of the dual-lens variety. This is shown in Fig. 2-20. Element K is the electron-emitting cathode surface. Surrounding this element is the control grid, CG, which has an opening facing the emitting surface. Further along, and coaxial with the control grid electrode, is the first, or focusing anode A_1; next in sequence, likewise coaxial, is the second, or accelerating anode A_2. By application of the proper operating voltages, the control grid is made negative with respect to the cathode, and likewise with respect to anode 1. This anode, on the other hand, is negative with respect to anode 2, since the voltage on the latter electrode is higher, relative to a common reference point.

Such a system, while not duplicated exactly in every electrostatic-focusing electron gun, is sufficiently typical to be usable as an example. Two electron lenses are formed by this assembly. The first lens, or cathode lens, consists of the field between the control grid and anode 1; whereas, the second lens is formed by the field existing between the surfaces of anodes 1 and 2. The similarity between the structure of the first lens, especially the bulging of the field through the apertures, and the parallel disks with holes in them (shown in Fig. 2-16), is very marked.

Inasmuch as this discussion of electron guns describes only the focusing action involved, the present explanation will be restricted to generalities concerning those principles. Accordingly, the first lens — the cathode lens — performs a specific function. It may be called the first focusing lens, because it tends to focus the bundle of electrons which pass through the grid aperture, but its *main function* is to produce a bundle of electron rays of minimum diameter, which becomes the object of focusing the system. *This is the first cross-over point and it is this point which is finally imaged on the screen.* Whatever may be the dimensions of the first cross-over point, as determined by the action of the control grid and the first anode voltage, those are the dimensions of the beam when it is properly focused by the second lens and strikes the screen.

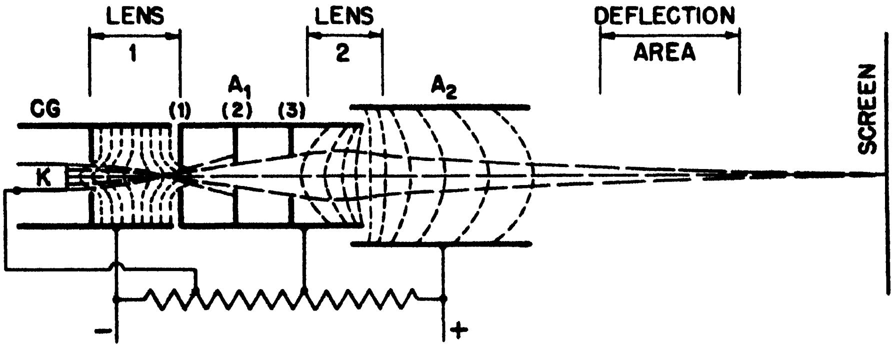


FIG. 2-20. — Practical electrostatic focusing system used in cathode-ray tubes.

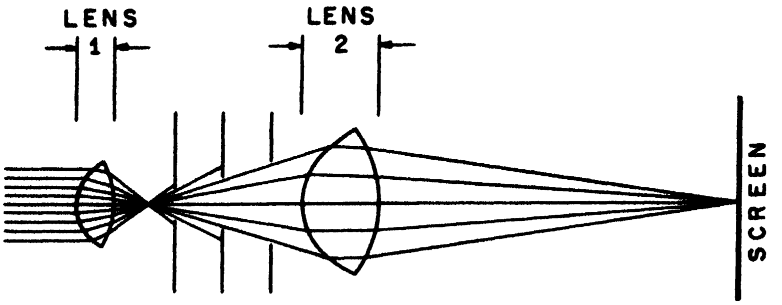


FIG. 2-21.—Optical equivalent of electrostatic focusing system show in Fig. 2-20.

Based upon what has been said, it is not difficult to see that any condition which will alter the distribution of the field in the first lens, such as a change in the voltage applied to the control grid or to anode 1, will have an effect on the dimensions of the first cross-over point. It will also affect the angle of divergence at which the electron beam leaves the cross-over point and, to a limited extent, the position of the first cross-over[2]. This statement is not made to indicate that these voltages must remain static and once adjusted may not be changed. It is simply a statement of the action of this field, which, as you will see later, was productive of certain problems and caused such changes in gun design as tended to eliminate undesirable effects.

While it is assumed that the emitted electrons are formed into a tight bundle in the cathode lens, there are nevertheless fringe electrons, because all do not advance at the same velocity, and are not equally affected by the lens. Moreover, all of the electrons which are pulled through the control-grid aperture are not tightly packed. Finally, the operation of the first lens system, even under ideal conditions, may not produce a first cross-over point of the desired small diameter, so that baffles are contained in the electrodes of this lens to mask off the fringe electrons and to aid in the development of a dense beam of small cross-sectional area.

Further control of the beam width takes place within the confines of the first anode through the use of the apertured baffles, *2* and *3*. This area, however, is considered to be free of field, so that whatever the path of the electrons may be, as they pass the vicinity of the first aperture *1* in anode 1, that path is maintained and they travel in diverging lines until they reach the principal focusing region, or lens 2. This is the space where the anode 1 and anode 2 cylinders face each other.

Here the diverging paths of the electrons continue for a short distance and then change to converging paths as they enter the right-hand half of the field. The curvature of the equipotential surfaces in this lens is strictly a function of the ratio of voltages applied to these two cylinders, and a change in any one of these will influence the focusing. Since anode 1 is operated at the lower voltage, it is the one which generally is subjected to a change in voltage in order to alter the focal length of this lens. It must be understood that the focal length of this lens is variable, and when a change is made, it is occasioned by one of two conditions. Either the position of the first cross-over point has changed, or some variation has occurred in the potential distribution of the second lens. As long as the ratio of voltages remains the same on anodes 1 and 2, in any one system, its behavior as a focusing lens will remain unchanged. It is not the specific values of the voltages applied which develop certain contours for the equipotential surfaces, but rather the *ratio* of the voltages applied to the two coaxially symmetrical cylinders. A given ratio of diameters d_2/d_1 for two such cylinders will, for given voltages V_2 and V_1, always produce the same distribution of potential[3]. Under the circumstances, the behavior of such a device as an electron lens always will be the same, since V_2/V_1 can equal V'_2/V'_1 regardless of what the individual values of these voltages may be.

The optical equivalent of such a typical electrostatic-focusing electron gun is shown in Fig. 2-21. Very little comment is necessary since the similarities of the two systems are quite evident. It might be well to mention, however, that contrary to what might be found in a projection system, the point source of light which is imaged on the screen is the cross-over point between the lenses, rather than the original source of light whose rays are incident upon lens 1. If we imagine the principal focusing lens 2 maintained constant in position, a change in the location of the first cross-over point will alter the point where the rays issuing from lens 2 will meet. In turn, if anything is done to alter the behavior of lens 2, that is, change its refractive index, the meeting point of the rays issuing from that lens may be ahead of or in back of the screen, rather than at the screen. In either of the first two cases, the image will not be accurately that of the object — it will be improperly focused.

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[2] Zworykin, V. K., Morton, G. A. *et al.* "Electron Optics and the Electron Microscope," pp. 33-37, John Wiley and Sons, Inc., New York, N. Y., 1945.
Soller, T., Starr, M. A., and Valley, G. E., Jr., "Cathode Ray Tube Displays," 1st ed., p. 46, McGraw-Hill Book Co., New York, N. Y., 1948.
Maloff, I. F. and Epstein, D. W., "Theory of Electron Gun," *Proc. I.R.E.*, pp. 1386-1411, vol. 22, 1934.

[3] Epstein, D. W., "Electron Optical Systems of Two Cylinders as Applied to Cathode-Ray Tubes," *Proc. I.R.E.*, pp. 1095-1139, vol. 24, 1936.

Review of Electrostatic Focusing

For the individual who may be interested in a somewhat more rigorous approach to the basis of electron lenses, the refraction of electrons in an electric field, the following is offered. It may be omitted by the person unfamiliar with mathematics without impairing the usefulness of the material which preceded it. An analysis based on the behavior of a single electron is valid for a beam of such charges. Consider for the moment the illustration shown in Fig. 2-22.

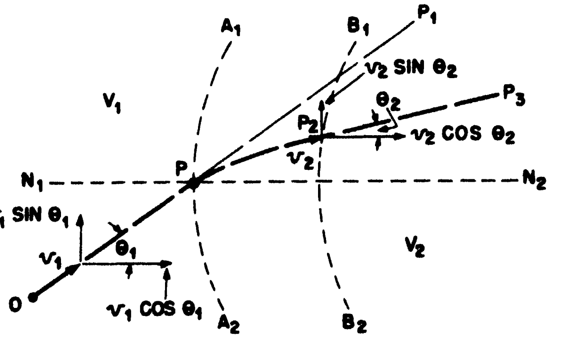

FIG. 2-22.—Refraction of electron within electrostatic field bounded by A_1-A_2 and B_1-B_2.

A_1-A_2 and B_1-B_2 are two equipotential boundaries, separated by a small distance. The region to the left of A_1-A_2 is theoretically field-free (no lines of force), although at a potential of V_1. The region to the right of B_1-B_2 is likewise assumed to be field-free, but at a potential of V_2. The *potential gradient* is in the direction of V_2; in other words, V_2 is more positive than V_1, and both are positive with respect to some point from which an electron, initially at rest, has started.

The path of the electron as it advances toward A_1-A_2 is along O-P, where P is the point at which the electron crosses the equipotential surface A_1-A_2. The line N_1-N_2 is normal to this surface at point P, and is also the direction of the electrostatic line of force crossing the surface at this point. The velocity V_1 of the electron as it crosses A_1-A_2 at point P may be resolved into two components — a transverse component, $v_1 \sin \theta_1$, and a longitudinal component, $v_1 \cos \theta_1$. The former is perpendicular to the field normal, while the latter is parallel to the normal. Upon entering the field between the two equipotential boundaries, the charge will tend to advance in the direction of the field — that is, parallel to the field normal. This results in an increase in longitudinal velocity, now identified as $v_2 \cos \theta_2$. The transverse velocity, $v_2 \sin \theta_2$, remains the same as before. A new electron path, P-P_2, is, therefore, established within the space, this path being curved toward the normal, instead of the straight path P-P_1. The electron then crosses the second equipotential boundary surface, B_1-B_2 into a field-free area and continues along the straight-line path, P_2-P_3. This line is a continuation of the path of the electron as it crossed boundary B_1-B_2 into the field-free area.

The curving path taken by the electron in the space between the equipotential boundaries shows a difference between the behavior of an electron lens and an optical lens. In the optical system, the refracted path of the light ray is a straight line, caused by abrupt changes in the refractive indices of the transmitting media. In the electron lens, the refracted path of the electron is curved because there is a progressive change in the refractive index; this is due to the infinite number of equipotential surfaces in a given field (in this case, between A_1-A_2 and B_1-B_2). Consequently, the horizontal component of velocity increases gradually, as each equipotential surface is crossed.

The kinetic energy of a moving electron is proportional to the potential difference, so in the region V_1 it may be expressed as

$$\tfrac{1}{2}mv^2 = eV, \tag{2-4}$$

therefore

$$v_1 = \frac{\sqrt{2eV_1}}{m} \tag{2-5}$$

and

$$v_2 = \frac{\sqrt{2eV_2}}{m} \tag{2-6}$$

The vertical component of velocity remains constant, thus

$$v_1 \sin \theta_1 = v_2 \sin \theta_2 \tag{2-7}$$

and, substituting from equations (2-5) and (2-6),

$$\frac{v_2}{v_1} = \frac{\sin \theta_1}{\sin \theta_2} = \frac{\sqrt{V_1}}{\sqrt{V_2}}. \tag{2-8}$$

Snell's law of optics for the path of a light ray traveling from one region of refractive index n_1 to another of index n_2 is shown in Fig. 2-23, where i is the angle of incidence and r is the angle of refraction and

$$n_1 \sin i = n_2 \sin r. \tag{2-9}$$

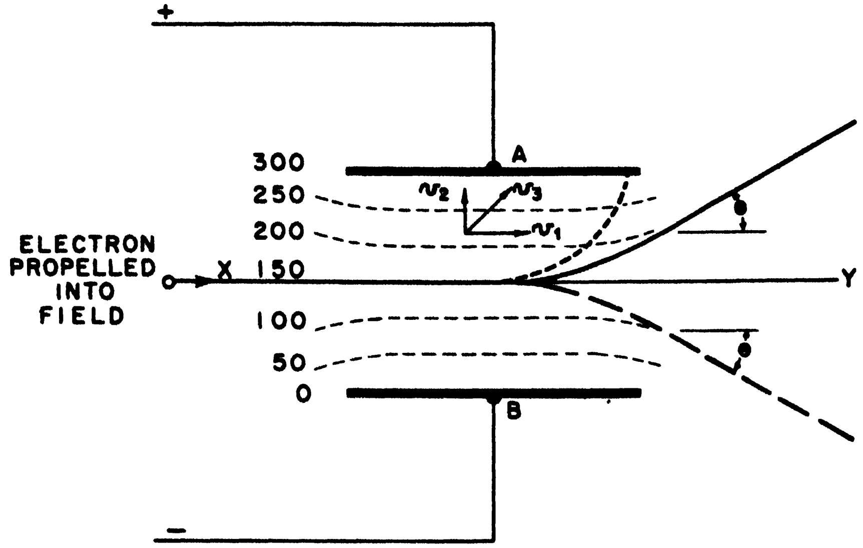

FIG. 2-23.—Deflection of electron between two parallel plates.

With angles θ_1 and θ_2 analagous to the angles of incidence and refraction in optics, and V_2 and V_1 being considered the equivalent of the refractive indices n_2 and n_1, the similarity between the laws of electron optics and light optics is evident.

In light optics, the greater the index of refraction of the medium through which the light ray is passing, the greater is the amount of bending of the light ray. In electron optics, the greater the increment of voltage between two points in the field being crossed by the electron, the greater will be the amount of bending of the path.

So much for the basis of electrostatic focusing. Although the physical structures of all electron guns which employ electrostatic focusing of the electron beam are not necessarily the same as the typical system illustrated in Fig. 2-20, the fundamental conditions of operation remain unaltered. The introduction of additional elements for further control of the electrons admitted into the focusing field, or the introduction of additional elements to increase the velocity of the electrons in the beam, does not change the basic laws of electrostatic focusing as described herein. No matter how the field is obtained, the basis of all electrostatic electron lenses is that a force acting on an electron, which is traveling through a field, is in a direction perpendicular to the surfaces of equal potential and is proportional to the intensity of the field at every point.

Electrostatic Deflection

In line with what was said earlier about the action of the electron within the uniform field between two parallel plates, it should not be difficult to visualize the process of electrostatic deflection. In the system shown in Fig. 2-23, which, incidentally, is a reasonably close approach to present usage in cathode-ray tubes, the two plates, *A* and *B*, are parallel to each other and form a deflecting structure. A source of voltage sets up a difference of potential between the two plates, which, for the sake of convenience, we will assume to be 300 volts, with plate *A* positive with respect to plate *B*. Since the plates are parallel, a uniform field is set up. This is so, regardless of the fact that both the field intensity and polarity may be temporary, and subject to a complete reversal or a change in value. All of this is incidental to the fundamental actions involved. An electron is propelled into the field along the axis *X-Y* at right angles to the lines of force, through which it is assumed to advance at a substantially constant velocity.

While in the field, the electron is experiencing a constant cross-wise force and, at the same time, it is advancing linearly through the field. The result is a *parabolic* path while within the field — a path which resembles the trajectory of a bullet fired from a gun. The curvature of the path is in the direction dictated by the direction of the lines of force (the cross-wise force), and after the electron leaves the field, it continues in a straight line, as shown in Fig. 2-23. The angle of deflection, θ, is that angle made between the path of the electron as it leaves the field and a line parallel to the axis of the system.

The Angle of Deflection

The angle of deflection, θ, is proportional to the length of the field, that is, the length of the deflecting plates, and to the ratio E_f/E_a, where E_f is the difference of potential between the plates and E_a is the accelerating voltage which propels the electron through the field. The basis for this relationship is easy to understand. For any one velocity of the electron as it is advancing longitudinally through the field, and for any one value of potential difference between the plates, the longer the plates, the longer is the time that the transverse field can act upon the electron so as to deflect it from its original path; consequently, the greater will be the final angle of deflection.

For any one dimension of the field and for any one longitudinal velocity of the electron through the field, the greater the potential difference between the plates, the higher the intensity of the field; consequently, the greater will be the force on the electron toward the positive plate while it is in the field. Since the velocity of the electron as it is passing through the field is substantially constant, the higher the potential difference between the plates, the greater will be the transverse displacement of the electron per unit time; therefore, the greater will be the angle θ.

Finally, the higher the longitudinal velocity of the electron through any one field of fixed dimension and of fixed difference of potential between plates, the greater will be the longitudinal displacement for a given amount of transverse displacement. The result is a decrease in the angle θ and in the deflection. Or, stated in another way, the higher the velocity of the electron, the shorter the time it will remain in the deflecting field; consequently, the deflection will be smaller.

The ultimate path of the electron through the field can be resolved into two components of motion, v_1, which is the longitudinal velocity along the axis, and v_2, which is the velocity in the transverse direction under the influence of the field. These two combine to produce v_3, the resultant path.

As was already mentioned, the direction of the deflection, which is indicated by the direction of curvature of the path, is a function of the direction of the field between the plates. The solid-line path in Fig. 2-23 is for plate A positive with respect to plate B, with the resultant deflection toward plate A. If, however, the polarities were reversed, making plate B positive, the direction of curvature of the path would be toward plate B, since it would now be positive. This is shown by the long-dashed line in Fig. 2-23. Assuming that the difference of potential between the plates is the same, regardless of the relative polarity of the plates, the angle of deflection, θ, is the same for the two cases illustrated. If, for some reason associated with the source of the voltage which establishes the difference of potential between the plates (the deflecting voltage), a higher difference of potential will exist with plate A positive with respect to plate B than when plate B is made positive with respect to plate A, then the deflection angle will be different for the two cases.

Since the deflection is proportional to the voltage applied across the plates, it is conceivable that the deflecting voltage may be so great as to bend the path sufficiently to cause the electron to strike the positive plate, thus preventing it from leaving the confines of the deflecting area. This is shown by the short-dashed line curve in Fig. 2-23. Obviously the same condition may exist for either relative polarity of the plates. As we have mentioned earlier, the motion of the electron due to the transverse field is the result of both attraction by the positively charged plate and repulsion by the negatively charged plate.

The representation of the deflecting action in Fig. 2-23 might be considered as a side view of deflection in the vertical plane — that is, upward and downward from the axis. Such orientation is strictly a matter of the orientation of the two deflecting plates. As seen, they are horizontal, but because the lines of force of the field always are perpendicular to the plane of the charged surfaces, and the deflection of the electron follows the lines of force, deflection of the electron is at right angles to the plane of the deflecting plates. Thus, deflection plates which are oriented *horizontally* with respect to the earth produce *vertical* deflections, and deflecting plates which are oriented *perpendicularly* with respect to the earth produce deflections in the *horizontal* direction. However, the same principles apply to both.

While it may be obvious, nevertheless it will do no harm to mention that if the difference of potential between the plates is zero, due either to the complete absence of any voltage, or because the same voltage is applied to both, the path of the electron will be a straight line along the axis X-Y.

Path of the Electron After Leaving the Field

The path of the electron after leaving the deflecting field is a straight line, tangent to the path at the point where the electron leaves the field. The absence of curvature after leaving the field results from the natural tendency of moving bodies to follow a straight path unless applied forces cause it to deviate. The transverse, or deflecting, component is now gone, and the electron advances unimpeded until it strikes a surface or object which prevents it from going farther.

It is not too much of a jump to imagine the single electron of Fig. 2-23 replaced by a stream of electrons, in which case the deflection of an electron beam will conform to all of the conditions stated for the single electron.

CHAPTER 3

PRINCIPLES OF ELECTROMAGNETIC DEFLECTION AND FOCUSING

The fundamental phenomenon underlying the focusing of electrons into a beam by means of electromagnetic fields is the reaction between two magnetic fields. This may not be evident from a visual analysis of the conditions of operation within the cathode-ray tube, but it is nevertheless true. One of these magnetic fields stems from the motion imparted to the emitted electrons which eventually form the electron beam. The creation of such a magnetic field is a basic law of physics. The other field likewise is a natural phenomenon, but its point of origin, physically speaking, is different, being the current which is made to flow through a focusing coil. The magnetic field within the core of this coil reacts upon the magnetic field surrounding the emitted electrons, which are made to pass through the opening in the core of the focusing coil. An idea of this relationship between the two elements of the system is shown in Fig. 3-1.

The electrons are emitted from the gun which contains the cathode, the control grid, the screen grid, and the accelerating anode. After emission and partial concentration of the beam by the creation of the first crossover point, in the neighborhood of the control-grid opening, as previously described in connection with electrostatic focusing, the electrons are accelerated toward the deflection area by the application of a positive voltage on the accelerating anode. This element is hollow, and the electron stream passes through it. The electromagnetic focusing coil also is hollow and fits over the neck of the glass envelope. It is so positioned, relative to the electrodes within the tube, that the core area of the focusing coil approximately embraces the space, or a portion of the space, occupied by the accelerating anode.

The focusing voltage is applied to the focusing coil and the current in the winding establishes a uniform field within the core space. Since the magnetic field can penetrate the glass envelope, as well as the nonmagnetic metallic substances of which the accelerating anode is made, the entire space embraced by the core of the coil is filled with magnetic lines of force. It is through this field that the emitted stream of electrons is pulled by the accelerating anode. Since the moving electrons are surrounded by a magnetic field of their own, the core space is filled by two magnetic fields, that due to the current through the focusing coil and that due to the moving electron stream. Magnetic lines of force *cannot cross each other,* because if they did, the field at the cross-over point would have two directions; hence, the two fields interact and create just one field, which represents the combination of forces present in the two fields, and which is called a resultant field.

The result of this interaction is control of the electrons in the stream, such that they follow a prescribed path established by the magnetic lines of force. As you will see later, this is a helical, or corkscrew-shaped path, with the ultimate effect that the comparatively widely separated electrons in the stream are concentrated into a beam as they leave the focusing coil and enter the deflection area.

It might be well to mention in passing that electrostatic focusing is not entirely absent in the electromagnetically focused tube. We include this reference at this time, only because electrostatic action is implied

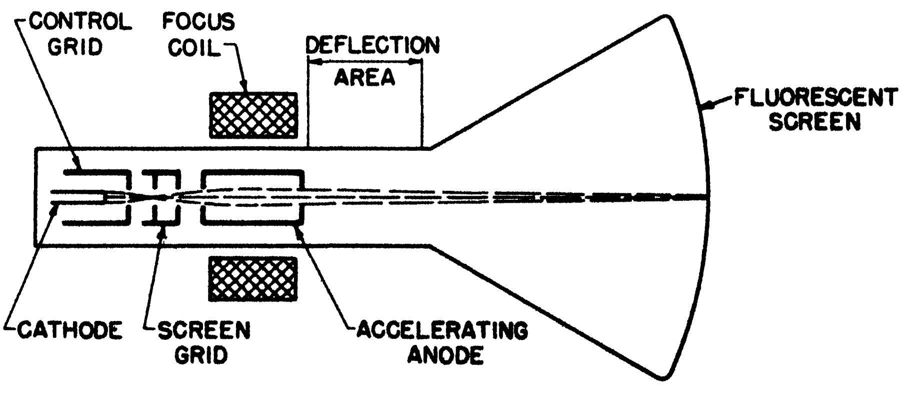

FIG. 3-1. — Relative positions of internal elements and external focus coil of electromagnetic cathode-ray tube.

when we speak about a control-grid element as well as an accelerating anode being within the electron gun — especially since these are at different potentials. Previously, we stressed the existence of an electrostatic field between any two points which are at different potentials; therefore, it stands to reason that in this case, too, there must be such a field.

The contribution of this field, or for that matter the electrostatic field which exists in all tube structures where a screen-grid element is located between the control grid and the accelerating anode, is not great. The principal focusing is attained by means of the focusing coil; nevertheless, we must recognize the existence of the electric field. We might say that these electrostatic actions, which precede the electromagnetic focusing, make the latter easier to accomplish by tending to confine the electron stream, as it comes within the influence of the magnetic field.

As can be readily realized, the foregoing is a brief description of the over-all action, not an explanation of how the result is accomplished. It might be a satisfactory presentation to the individual with only a casual interest in cathode-ray tubes, but the person who will use the cathode-ray tube as a daily tool desires a more detailed analysis.

Electromagnetic versus Electrostatic Focusing

Both electrostatic and electromagnetic focusing methods accomplish the same end. However, while the results are the same, the methods differ. This may lead to the question, "Why electromagnetic focusing, if electrostatic focusing accomplishes the same thing?" The answer has many facets, a few of which shall be dealt with in brief. To begin with, many things can be accomplished in two ways, so that the existence of two methods of focusing is in itself not surprising. Moreover, each of these methods has its advantages and disadvantages and it is simply a case of utilizing that which best fits the need.

Electromagnetic focusing affords the advantage of simpler tube construction. If you will compare the elementary presentations of the electron guns shown in Figs. 2-2 and 2-3, you will note that the gun used in the magnetic type of tube has fewer elements; the multiple focusing electrodes used in the electrostatic gun are omitted. Of course, the magnetic tube requires a focusing coil, which is not necessary in the electrostatic type of tube, so that one might seem to offset the saving in the other.

Simplicity of tube construction, or the relative cost of the focusing electrodes in one case and the focusing coil in the other, are not the only factors to be considered. The electrical factors relating to voltage and current also require attention. It is generally said that "current is cheaper than voltage", by which is meant that the cost of the device which will provide a high voltage at low current is generally greater than the cost of the device which will provide a comparatively high value of current at low voltage. Moreover, there are times when the control of a high voltage is fairly difficult. Evaluating all of the factors, we find that electrostatic focusing is preferable for one class of cathode-ray tube, whereas electromagnetic focusing is preferable for another. The former is the usual run of comparatively low-voltage cathode-ray tubes used in laboratory and maintenance applications as well as the small-screen television receivers. On the other hand, electromagnetic focusing is found preferable for the high-intensity type of cathode-ray tubes where very high voltages are required for acceleration in order that the image on the screen be brilliant. In such cases, very high values of focusing voltage would be needed, and the control of these would be more difficult than the control of a high-current low-voltage focusing system.

An advantage of the electromagnetic focusing type of tube is the reduced defocusing effect caused by the deflection process. This does not imply that all electrostatic tubes suffer from this defect, for it is well known that sharply focused images are usually obtained with electrostatically focused tubes. As a matter of fact, both electromagnetic and electrostatically focused tubes suffer some defocusing in the deflection process; the effect is generally greater in the case of electrostatic focusing, but even then it is not always noticeable. Much of the correction is taken care of in the design of the tube.

The line of demarcation between the different classes of service to which these two focusing methods are put is not as fine as might be gathered from what has been said. We have generalized, and it is not surprising that both types of focusing will be found in systems which make use of tubes of different dimensions for different purposes.

To a limited extent, the type of deflection has a bearing upon the type of focusing used; that is, a tube which is intended for electromagnetic deflection usually will be of the electromagnetic focusing variety. We say usually, because it is not rare to find systems making use of tubes which have electrostatic focusing and electromagnetic deflection.

One condition bears repetition, however, namely that laboratory and maintenance types of cathode-ray-tube equipments invariably have electrostatic focusing and deflection.

To explain how electromagnetic focusing is accomplished in the cathode-ray tube, it might be well to begin at the elemental charge. This does not mean that we shall review every subject included in the general category of elementary electricity and magnetism. Such is not necessary, since the individual who has occasion to use the cathode-ray tube will have advanced sufficiently in his studies of these subjects to be acquainted with the elementary facts. All we shall deal with are those facts which are pertinent to the attainment of our goal — an understanding of electromagnetic focusing and deflection.

The Moving Charge

When we examined electrostatic focusing in the preceding chapter, we made clear that there existed an electrostatic or electric field around every individual charge. This field was shown to be an attribute of the charge; wherever the charge existed, the field existed. Where equal numbers of unlike charges existed, the net *external* effect of the electric field around each charge was zero, but the fields nevertheless existed. Where a preponderance of any one kind of charge existed, the field existed *between* these charges and an equal number of opposite charges, which may have been close by or at a remote point.

As it happened, we concerned ourselves at that time principally with *charges at rest,* or at least *moving so slowly* that they could be considered as being at rest. Whatever motion was associated with the redistribution of charges, so as to establish a net electric field between any two points, was viewed as being strictly temporary and a means to an end. It is, however, conceivable that failure to associate the moving charge with any other related phenomenon may have created the impression that only one kind of field is, and can be, associated with the elemental charge — this being the electrostatic field when the charge is at rest.

If such an impression does exist, a correction is required, because the elemental charge may be associated with two types of fields. When it is at rest, the only field existing around it is the electric field. When it is in motion, two fields exist around it — the electric field, and a *magnetic field which has come into being as a consequence of the motion of the charge.* If such motion ceases, the magnetic field disappears. But whether at rest or in motion, the electric field is always associated with the charge.

Two general classes of motion of the charge, and their corresponding magnetic fields, should here be mentioned. The charge, or charges, may move at a steady rate, in which case a constant magnetic field is developed; or the charges may move at a varying velocity, or alternate back and forth, which will produce a changing or alternating field. If the character of the motion repeats itself at regular intervals, the motion and the resulting field are said to be *periodic.* It is quite possible that a combination of a steady and an alternating current and field may occur; in most cases, the effects of the two components may be considered separately. Both types of fields find application in cathode-ray tubes and will be dealt with at the appropriate places in this text.

The following is a condensed tabulation of the effects of electric and magnetic fields:

Stationary Electric Field	*Stationary Magnetic Field*	*Moving (AC) Electric Field*	*Moving (AC) Magnetic Field*
Will act on stationary electron.	Will not act on stationary electron.	Since each is an attribute of the other, the two act on either stationary electrons or electrons in motion, and will act upon a magnet.	
Will act on electron in motion.	Will act on electron in motion.		
Will not act on magnet.	Will act on magnet.		

The relationship among all the conditions pertaining to the electric and magnetic fields, around a charge in motion, is not expressed fully by these relatively few statements. Much more can be said, but since it is not necessary to do so at this time, we restrict the extent of the discussion in order to most effectively associate the pertinent ideas.

If you are wondering why we did not mention the existence of the magnetic field during the explanation of electrostatic focusing, there is a simple explanation. The magnetic field plays no part in connection with electrostatic phenomena; therefore, it did not justify reference. It is like speaking about magnetic fields when analyzing a filter system. The charges which are in motion within the elements of the filter are surrounded by a magnetic field, but the importance of this field with respect to the subject at hand is negligible; consequently, any mention of the magnetic effect while explaining filter action would, to say the least, needlessly confuse the issue. The same applies to the comments concerning a magnetic field during the explanation of electrostatic focusing — since it would have added nothing of value to the discussion, it was omitted.

Now the situation is different. We are interested in electromagnetic focusing and the magnetic field is important. On the other hand, the electric field is of little consequence, and therefore, shall not be treated during the discussion of the magnetic effects of the moving charge, or current.

The Electromagnetic Effect

That there is some effect when the electron moves, which is nonexistent when the electron is stationary, was discovered more than 120 years ago by the Danish physicist, Oersted. At the time, he had no conception of electrons. However, he discovered that when current was flowing through a wire which was in the proximity of a compass, certain conditions came into being which were capable of exerting a force upon the compass needle, so that its orientation was changed from that normally dictated by the existence of the earth's magnetic field. He found that changing the direction of the current through the wire, by reversing the terminals of the voltage source, reoriented the compass needle; see Fig. 3-2.

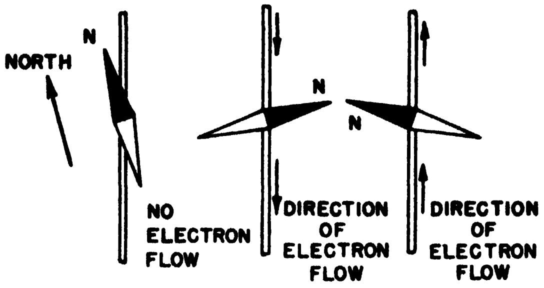

FIG. 3-2.—Oersted's experiment illustrating effect of magnetic field around current-carrying conductor.

In both cases, he observed a very interesting condition; the compass needle tended to turn so that it was perpendicular to the wire, as shown in Fig. 3-2. Moreover, he noted that whatever was the phenomenon under observation, it had something to do with the poles of the compass needle, because the north pole was urged in the opposite direction when the direction of the current was changed. The exact nature of the space around this current-carrying wire was not known, but it is not strange that it was recognized as a special condition akin to magnetism.

By that time, many things relating to magnets and magnetism were known, although naturally not as much as developed later; but the concept of poles had been advanced hundreds of years before. And since a similar rotational movement of the compass needle was possible by means of forces due to other magnets, it is not unreasonable that some association between electric current and magnetic effects should be suspected. Identification of the condition of space about a current-carrying wire as a magnetic field, and its mapping by lines of force by Faraday, was yet to come. Whatever it was that the electric current brought into being, its effects were very much like those of the magnet. The discovery revealed that electric current could produce magnetic effects!

The Magnetic Field

Today we know that a wire which is carrying electric current is surrounded by a condition of space which the physicist calls a magnetic field; the name no doubt is the result of an association of effects, those of the current being the same as those of a magnet. It is an area in which forces exist, manifesting themselves in the same manner as the forces which exist around a magnet.

If we view current as being charges in motion, then, as we said before, the electron in motion is surrounded by a magnetic field. Therefore, *every segment* of a wire in which current exists is surrounded by a magnetic field, because every segment of the wire has in it charges which are going through directed motion. However, it is very important to understand that the presence of a physical path — the wire — is incidental to the magnetic field which exists around the moving charge. It would exist just as readily if there were no physical conducting path, even if charges were moving through a vacuum. Since physical paths for electron movement are commonplace, it might be well to assign the term "electron flow" to the directed motion of electrons in contra-distinction to the random motion of electrons in such a physical medium when it is not carrying a current[1].

Like an electrostatic field, the magnetic field is intangible. It cannot be seen, and its presence is detected only by the effects it produces. Science has seen fit to develop a concept of the organization of such a field in terms of magnetic lines of force. Actually this concept was developed by Faraday, who envisioned the magnetic forces acting along imaginary "lines of force". If the field around a current-carrying conductor were mapped, it would resemble the pattern shown in Fig. 3-3A.

Some very interesting and important observations can be made about these lines of magnetic force. You will recall that electrostatic lines of force were single lines which joined unlike charges. These lines started from and terminated upon charges. Magnetic lines of force, on the other hand, are not attached to the charge;

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[1] The familiar "arrow" convention for the direction of electron flow, looking at the end of a conductor, is used in this book. Thus if the electron flow is out of the page, it will be indicated by a dot in the center of the conductor, representing the head of the arrow. Conversely, if electron flow is into the page, a cross, representing the tail-feathers of an arrow, will be placed in the conductor.

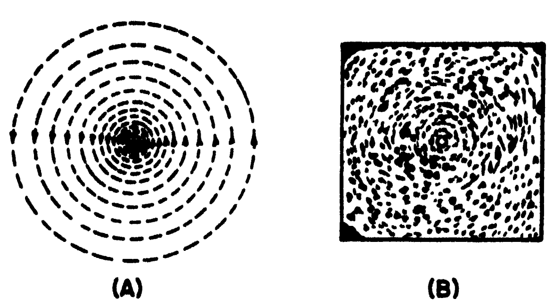


FIG. 3-3.—Idealized concentric field around current-carrying conductor (A), and actual effect on iron filings (B).

they surround the charge as it moves, with the charge occupying the center of each loop. In other words, magnetic lines of force are *complete loops*, and whatever magnetic force is felt in a magnetic field, is felt along a closed *loop of force*.

Admittedly, a physical visualization of such loops of force is difficult, especially when we attempt to state Faraday's concept, wherein he viewed these loops, like the electrostatic lines, as behaving similar to stretched rubber bands, tending to make themselves as short as possible. The development of this concept, however, is secondary to the appreciation of the behavior of the magnetic loops, i.e., the conditions within such a field. A physical impression is not vital to the understanding of the actions, just as a physical impression of energy, which is invisible, is not essential to an appreciation of its meaning and uses. So we simply accept the condition that the space around a current-carrying wire, or around each moving charge, is filled with something which is capable of exerting a magnetic force, and, if plotted in direction, will *appear as a group of closed, or continuous, loops*.

Understandably, Fig. 3-3A shows only part of the truth. In the first place, the loops of force around a current theoretically extend to infinity, although their presence is detectable only within comparatively short distances from their origin, because their intensity decreases very rapidly. Moreover, we show what is the equivalent of the loops surrounding just a single segment of the wire. Actually, such loops surround all the elements of the current, forming what might be considered a series of concentric cylinders around the current. There are an infinite number of these cylinders surrounding any current-carrying conductor. However, the limitations in drawing an infinite number of lines result in spaces between these lines, as seen in Fig. 3-3A. While such spaces do not actually exist, they may be conveniently attributed to the fact that, like electrostatic lines of force, magnetic lines of force which have similar direction tend to repel each other, whereas those which have unlike direction tend to attract each other.

The pattern formed by the lines of force can be made visible by a simple experiment, often performed in school laboratories. A wire is placed vertically through a level sheet of cardboard, and a heavy current is passed through the wire. Iron filings are then scattered evenly over the cardboard. Inspection of the filings shows two things: first, that there is a tendency for them to fall near the wire, indicating that the magnetic field is more concentrated there than it is further away; and second, that the individual filings fall so that their lengthwise dimensions lie on circles concentric with the wire. Fig. 3-3B shows the results of the experiment, in which the circular pattern is unmistakable. If a compass is used to explore the region around the wire, the fact that the compass has a north pole and a south pole can be utilized to determine whether or not there is a specific direction associated with the field.

Direction of the Magnetic Field

Every magnetic field has a direction of action associated with each point in the field. Since it is impossible to conceive of a directionless force, if we remember that a magnetic field is a condition of space such that a force is exerted upon a magnetically active body, that force must have some direction. It is accepted practice to consider the forward direction of the force to be the direction in which a north magnetic pole tends to move.

As in the case of the electrostatic field, the drawings of a magnetic field are simply maps showing only a few lines of force which indicate, at any point, the direction of magnetic force at that point. The strength, or intensity, of the field can be judged by the closeness of the lines of force; the further apart they are, the weaker the field.

In the case of the field around the conductor, the fact that the loops or lines of force are closed is quite evident from the arrangement of the filings.

The rule for determining the direction is an easy one. Grasp the wire with the left hand, the thumb pointing in the direction of electron motion, that is, toward the positive terminal of the current supply; the fingers will then point in the direction of the concentric field surrounding the wire.

Looking at Fig. 3-4A, let us examine the orientation of a number of very small compass needles, imagined to be placed near the conductor. Eight points of indication are selected. With respect to the line of travel of the moving charge, the magnetic loop has a constant direction of rotation around the line, but with respect to the direction at any one point, every other point on the loop has a different direction. If we select two points, in this case two opposite compass needles, it is seen that they point in *opposite* directions. The direction of action on the left side of the charge is the exact opposite of that on the right side of the charge. The direction of motion of the charge will naturally change the direction of action of the field as a whole, but will not alter the relation *that opposite sides of the field will exert forces in opposite directions.* This will become clear when we compare Figs. 3-4A and 3-4B. Just how this condition is used will become evident as the text unfolds.

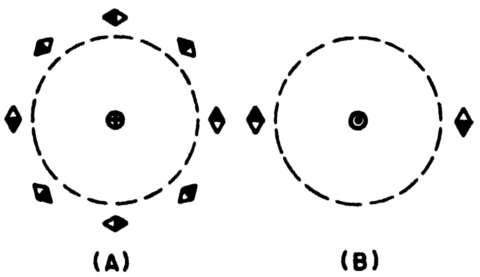


Fig. 3-4.—Effect of magnetic field, due to current-carrying conductor, on a number of test compass needles.

The final item of importance relative to direction of action of the magnetic field around the moving charge is the general condition that the plane of the field is at right angles to the direction of motion of the charge. In the instance of the Oersted experiment, the conclusion was that each loop of magnetic force lies in a plane at right angles to the direction of the wire. Changing the direction of the current may have changed the direction of action of the loop of flux, but did not change its plane with respect to the direction of the conductor. Since we are speaking about current as the moving charge, we associate direction with the motion of the charge, rather than the direction of the conductor.

Thus, three factors of direction affect the field around a charge. These are: (1) the direction of motion of the charge; (2) the direction of the line joining any selected point in the field to the path of the charge; and (3) the direction of the force at the selected point. All of these are mutually perpendicular, no matter what point in the field is selected. This is expressible in another way, viz; the plane of the magnetic field is at right angles to the direction of movement of the charge, while the direction of the force exerted at any point along the magnetic line is tangential to the line at that point. This is shown in Fig. 3-5, which is taken, with some modifications, from Fig. 3-3.

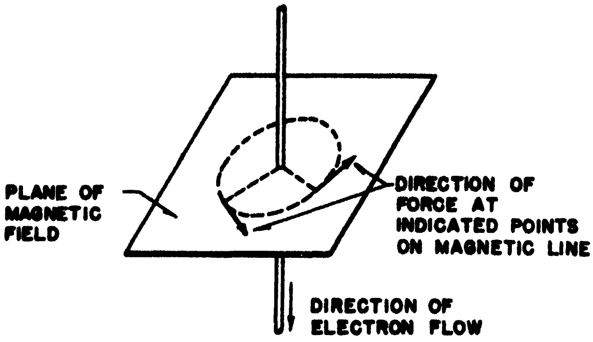

PLANE OF MAGNETIC FIELD
DIRECTION OF FORCE AT INDICATED POINTS ON MAGNETIC LINE
DIRECTION OF ELECTRON FLOW

Fig. 3-5.—Relationship of direction of current flow, plane of magnetic field, and direction of force due to magnetic field.

Arrows have been drawn, indicating the relationship described above. The dotted circle is included simply to show that the same condition exists all around the conductor.

Distribution of the Lines of Force

The distribution of the lines of force in the magnetic field is of special interest to us because in electromagnetic focusing, every effort is made to produce a special type of field, known as a *uniform field.* A uniform field is one wherein the intensity is the same at all points, or at least over that portion of space where the field is utilized. A further definition of a uniform field is one wherein the number of lines of force which thread a unit area, of say one square centimeter, will be the same everywhere in the field.

At first glance, one might imagine that a uniform field surrounds a single straight current-carrying conductor, simply because of the simplicity of the field. To one familiar with the convention in which field strength is indicated by the density with which lines of force are shown upon a drawing, it is immediately evident that the field shown in Fig. 3-3 is anything but uniform. The greatest concentration of lines of force exists in the immediate neighborhood of the conductor, and the density decreases as the distance from the conductor is increased. This indicates that the field strength is greatest close to the conductor and decreases as one moves away from the conductor. The field strength, H, in the case of such a single conductor is expressed by the Biot-Savart law as follows:

$$H = \frac{2I}{10r} \text{ oersteds} \qquad (3\text{-}1)$$

where I is the current in the conductor in amperes and r is the distance, in centimeters, between the center of the conductor and the point of measurement.

The same field strength exists at corresponding distances from the wire on all sides, provided that the angle, or orientation, of the plane of the test area with respect to the plane of the field, is the same. At all such locations, the same number of lines of force would thread the area. This is shown in Fig. 3-6A, in which two test areas, *A* and *B,* are equidistant from the conductor on opposite sides of the field. In the illustration, three lines of force thread each area. This number is purely illustrative and has no actual significance with respect to the magnitude of current in the wire. In Fig. 3-6B we have changed the position of test area *B*. The fact that it is on the same side of the field as test area *A* has no meaning; the important condition is that in being located closer to the conductor, more lines of force thread through it than thread through test area *A*. This indicates the nonuniformity of the field.

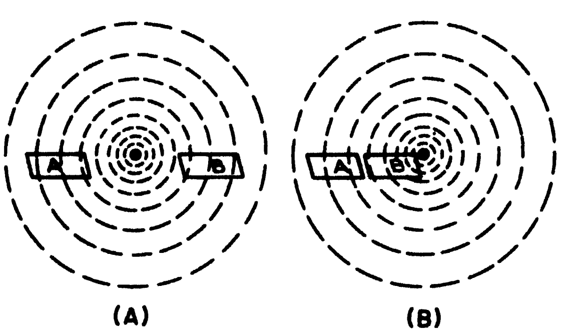


FIG. 3-6.—Nonuniform magnetic field around current-carrying conductor. Number of lines of force cutting identical test areas are equal in (A), unequal in (B).

The significance of such a condition is manifold, although for our purpose we need consider but a single effect, this being the difference in magnitude of force at different positions in a field of this type. Obviously, if some object susceptible to magnetic forces were placed at different points in the field, it would feel forces of varying magnitudes, and if this object were caused to move under the influence of these forces, the character of motion would be a variable determined by the changing amounts of force that it would experience as its position in the field changed.

How the electron behaves when projected into a magnetic field will be described later, but for the present you can accept the statement that if an electron were caused to move through the field so as to cut across the lines of force, the behavior of that electron would not be that which is desired, simply because the forces it would experience would not be constant in magnitude.

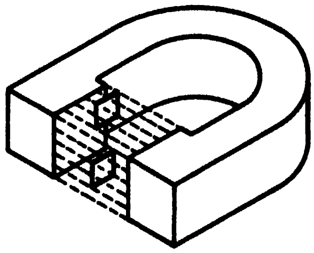


FIG. 3-7.—Uniform magnetic field between poles of horseshoe magnet. Same number of lines of force cut each test area.

As a comparison with what is shown in Figs. 3-6A and B, we illustrate in Fig. 3-7 the uniform magnetic field which exists between the poles of a horseshoe magnet. No matter where the unit test area may be located within this field, just so long as it is similarly oriented with respect to the lines of force, it will be threaded by equal numbers of lines. In view of the condition stated earlier, namely that electromagnetic focusing demands the action of a uniform field, the reference to a magnet may create the impression that permanent magnets are used for focusing. That is not correct. The aim is to produce a uniform field, but by making use of some form or configuration of a current-carrying conductor.

It is possible that the straight lines in Fig. 3-7 representing the magnetic lines of force and the loops shown in preceding illustrations may cause some confusion in your mind. There is no conflict between these two illustrations and what was said earlier about magnetic lines of force being complete loops. Such is the case, except that any very short segment of a circle can be viewed as a straight line, just as a small part of the earth's surface may be considered flat, while the whole earth is a globe. Each of the straight lines in Fig. 3-7 is really a part of a complete loop of force, the remainder of the loop threading its way through the magnet proper. In like manner, when we speak about lines of force threading the test areas in Figs. 3-6A and B, we are speaking about those *segments* of the loops of force which thread the area.

The Loop

If we change the configuration of the straight current-carrying conductor so that it is bent into a single-turn loop, a major change is made in the shape of the magnetic field which surrounds it. The concentric circles of force around the straight conductor now are changed to what might be described as elliptical loops, as shown in Fig. 3-8. Moreover, they become more crowded inside the loop than on the outside. Under the circumstances, the intensity of the field is greater in the space within the loop than in the space outside of the loop, for, as we said earlier, the intensity is a function of the number of lines of force which thread a unit area. If the lines are more crowded inside the loop,

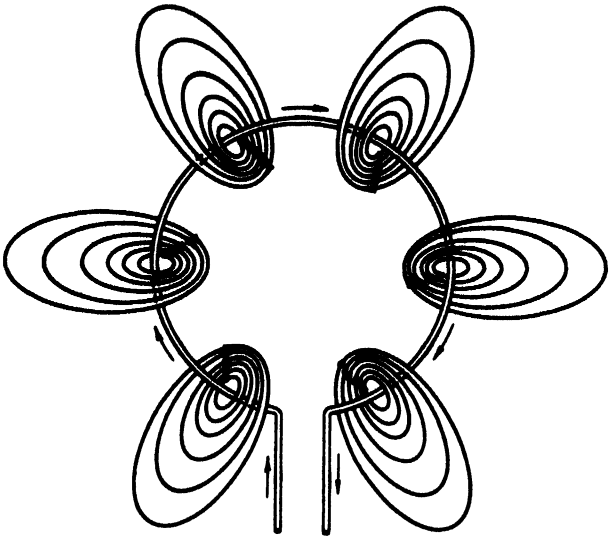


FIG. 3-8.—Magnetic field around single-turn current-carrying loop.

more of them will thread a unit test area than would thread the same unit area if it were located outside of the loop.

The magnitude of the field within such a loop of wire is expressed by the equation

$$H = \frac{2\pi nI}{10r} \text{ oersteds} \qquad (3\text{-}2)$$

where $\pi = 3.1416$ and $n =$ the number of turns, which, in this case, is equal to 1. The current I is in amperes, and r is the radius of the loop in centimeters.

It might be well if we stressed the fact that the portion of the field which is of interest to us is that located inside the loop. Since we must of necessity speak about what exists outside as well as inside the loop, the focal point of interest does not become evident unless it is pointed out.

In reference to the current-carrying loop shown in Fig. 3-8, several pertinent comments are necessary. First is the fact that the direction of each loop of force still follows the basic laws expressed in connection with the straight conductor—the direction of each loop is in accordance with the left-hand law. The second condition is that while we have shown loops of force *crossing each other, actually they do no such thing.* In this respect magnetic loops of force are like electric lines of force. They will react upon one another and produce a resultant field, but they will not cross each other, for if they did, the magnetic force at the point of intersection would have two directions, an impossible condition.

Coiling the straight wire into a loop has not changed the relationship between the planes of the conductor and the plane of the field. As can be seen, the two are at right angles to each other at any point, and the direction of the force associated with each line also remains, as before, tangential to the line. Finally, a single-turn current-carrying loop does not produce a uniform field, although as a consequence of limited space within the loop and virtually unlimited space outside the loop, greater concentration of the lines of force exists within the loop. Thus we have taken a step toward the attainment of the desired type of field.

The Solenoid

A further development of the uniform field, by means of a current-carrying conductor, is the use of more than a single turn in the coil. This leads to the *solenoid,* or coil consisting of numerous turns, either in a single layer or in many layers. An example of such a coil using a single layer is shown in Fig. 3-9. Each turn of the coil carries current, and therefore has a field of its own. By virtue of the positioning of the turns, these fields unite with each other and develop a resultant field through and around the solenoid. The pattern of the field is as shown, and corresponds in general to the pattern of the field existing around a bar magnet. In fact, such a solenoid is the equivalent of an electromagnet, in that it has its magnetic poles at each end, which are linked by loops of magnetic force. The magnetic polarity of the coil is determined by the direction of the current through the conductor, as follows: Grasp the coil with the left hand so that the fingers point in the direction of motion of electrons in the wire. The thumb then points to the north pole of the coil, where the loops of force leave it, to pass outside and re-enter the solenoid at the south pole. The similarity in magnetic field patterns around a solenoid and a bar magnet can be seen by comparing Figs. 3-9 and 3-10. In every respect, the electromagnet formed by the solenoid can be made to behave like the bar magnet and, while current is flowing through the coil, it will display all the properties of the bar magnet.

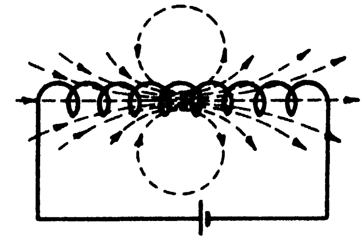


FIG. 3-9.—Magnetic field around single-layer solenoid.

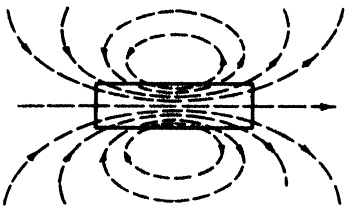


FIG. 3-10.—Magnetic field around bar magnet.

Many comments are possible concerning the field surrounding such a coil, but once again we are interested chiefly in the configuration of the field with respect to uniformity of distribution. In this respect the main point of interest is the condition inside the coil; in fact, it is this condition which dictates the use of many turns. We have in mind the *uniform field,* which is developed inside the core space of the coil. As in the case of the magnet, the path of each magnetic loop is partly through the magnet and partly through the space surrounding the magnet. The same is true of the solenoid; a part of the path of the loops of force is in the space within the coil, and the remainder is completed in the space outside the coil.

Although the field exists outside, its utility from the viewpoint of electromagnetic focusing is negligible. If anything, it may, under some conditions, be found desirable to confine it by magnetic shielding. However, insofar as this discussion is concerned, the field inside the coil is the important one. The greatest field uniformity exists at the center of the coil, and, as you can see, the lines are straight over a reasonably long portion of the coil length. Some curvature develops at the extreme ends of the coil, and it becomes a matter of proper coil design to make a uniform field available.

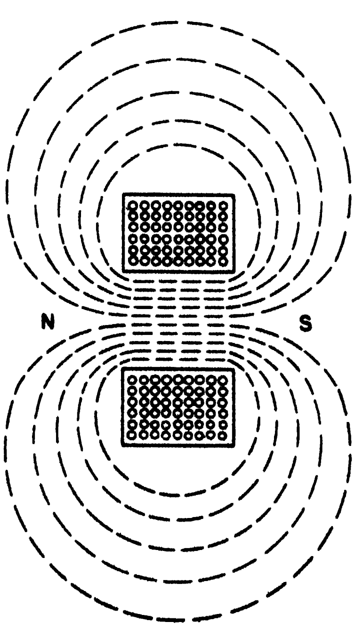


FIG. 3-11. — Approximate uniform field within core of multi-layer coil. Drawing shows coil in cross-section.

In the case of the single-layer coil, this calls for a high ratio of length to diameter.

Since coils which are thin and long may not be easy to handle, a modification of physical design is employed which makes for a more convenient winding, yet provides a uniform field with which to work. This is the multi-layer winding of sufficient length so that a substantially uniform field exists within an appreciable portion of the core length, as shown in Fig. 3-11. By utilizing the forces present over that portion of the field, within the core, which is made up of straight lines, the nonuniformity of the field in the neighborhood of the core openings can be neglected.

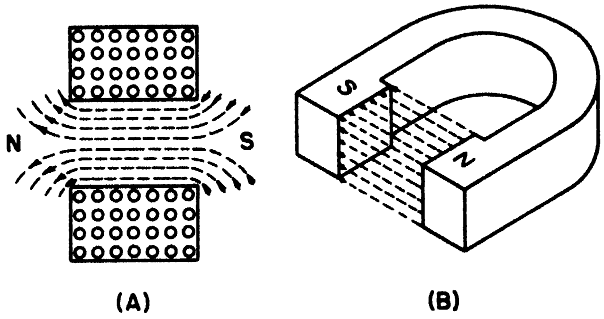


FIG. 3-12.—Uniform field within core of multi-layer coil (A) and horseshoe magnet (B). Direction of lines *within* core of coil is the same as direction of lines *within* magnet, but opposite to those in gap between poles.

Looking at the multi-layer coil as an electromagnet, it, like the single-layer coil, has its magnetic poles, and in this respect is no different from a conventional bar magnet. Therefore, the field within the core of the coil is the equivalent of the field which exists within the body of the bar magnet. We can stretch this equivalence a bit farther; assuming that we bend a bar magnet into a horseshoe shape, the field between the pole pieces corresponds to the field within the core of the coil, as shown in Figs. 3-12A and 3-12B. Since it is impossible to distinguish any difference in the character of magnetic lines of force issuing from a bar magnet or from an electromagnetic system, we can visualize the lines of force between the dotted lines in Fig. 3-12B as being those present between the south and north poles of a horseshoe magnet. In other words, the multi-layer coil gives a strong, uniform, magnetic field to work with.

Interaction Between Magnetic Fields

A few remarks have been made previously about the manner of interaction between magnetic fields. Before illustrating the mode of interaction, it might be well to say a few words about the behaviour of magnetic lines of force. These comments could have been made earlier, since examples of the resultant have been given, as in Figs. 3-9, 3-11, and 3-12, but they fit better here.

The magnetic loop of force might, in some respects, be compared to a stretched rubber band; the band, being under tension, tends to shorten itself. The same is true of the magnetic line of force, but this is about the limit to which we can go in making the comparison. From this point on, we must treat the loop of force by itself.

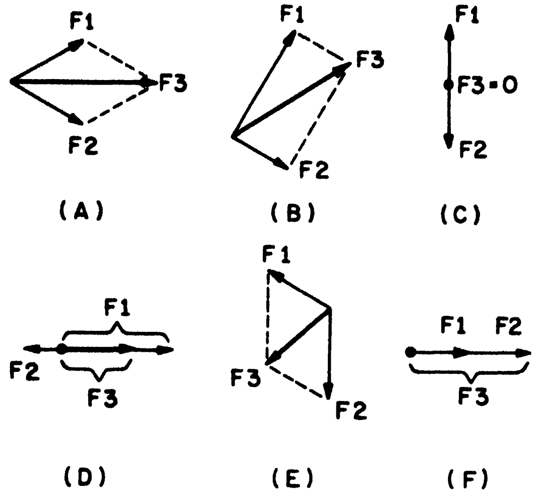


FIG. 3-13.—Component and resultant vector representations.

It is understandable that more than one force may exist at one time and, further, that each of these may be active in a different direction. For example, in Fig. 3-13 are shown several *vector* presentations of different forces, *F1* and *F2,* of different magnitude and direction, and the resultant force *F3.* In each case, the resultant has a direction and a magnitude which is determined by the individual directions and magnitudes of the component forces. Where the forces are equal in magnitude and exactly opposite in direction, as in Fig. 3-13C, the resultant force is zero. Where the forces are active in the same direction, the resultant is the sum of the two, and is obtained by making the origin of one vector coincident with the head of the other vector; the sum is the new vector which extends from the origin of the first vector to the head of the second vector, as shown in Fig. 3-13F.

It is possible in similar fashion to show, point by point, the resultant of two magnetic fields, inasmuch as every point in a field represents a certain magnitude as well as direction of force. In other words, points in a magnetic field may be stronger or weaker than other points in the same over-all field and will have a resultant direction which is determined by the components. At some points, the forces may aid each other, and at other points they may oppose each other. At every point, however, the *ultimate pattern of two fields,* which may be due to *two different currents,* will be the *resultant,* and there will be such rearrangement of the lines of force that at any one point in the field, each line has but one direction.

The rearrangement is, of course, a condition which is associated with the behavior of the lines of force. We said that they will tend to shorten themselves along their length, and, again like the electrostatic lines of force, a magnetic line of force will seem to repel another line of force which has the same direction. Herein lies the reason for the so-called greater separation between lines of force existing outside a loop or solenoid than prevails inside the loop. Moreover, the straight-line field within the core of a solenoid or a multi-layer coil is due to the mutual repulsion between the lines (all of which have the same direction) on all sides of the lines. It is the same condition which makes for the straight lines of electrostatic force between two large, closely spaced, parallel plates.

Obviously, where the forces in a field aid each other, the magnitude of the field intensity is greater than in areas where the forces tend to cancel or offset each other's effects. In accordance with common practice, illustration of such a condition is by means of the density of the lines of force.

FIG. 3-14. — Magnetic fields around conductors having current flow in same directions. Weakened field between conductors tends to force them together.

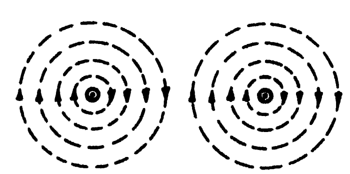


Fig. 3-14 shows the patterns of the resultant field around two conductors, in which the currents are flowing in the same direction (out of the page). In the case of the two wires which are carrying equal currents in the same direction, the direction of the magnetic loops is such that they are opposite on the sides which face each other and are advancing in a constant direction of rotation, on the two outsides. The net result is that the fields on the inside tend to offset each other, thus producing a weaker field between the wires, and effectively causing the fields on the outside to strengthen each other. Consequently, there is a concentration of lines of force on the outside with respect to the reduction of lines of force between the wires. This statement does not imply that more lines of force exist on the outside of the wire than between the wires; equal numbers of lines exist on both sides. It simply means that the intensity of the field is greater on the sides which are opposite (outside) than on the sides of the wire which face each other.

Such a condition gives rise to the physical effects of the force. In this case, that is, when the currents are in the same direction, the two wires tend to move toward each other. One explanation can be that the action is from the stronger field toward the weaker field. The field is stronger on the outside than between the wires; therefore, the wires move toward each other. A second explanation involves the previous statement that magnetic lines of force will tend to shorten themselves along their length. Such is possible if the two wires move toward each other. The third explanation is that any two circuits which are carrying current will tend to so position themselves that they will include the greatest number of lines of force common to both. This is accomplished if the two wires move closer together.

In the case of two wires which are carrying currents in opposite directions, the direction of the facing magnetic lines within the space between the wires is the same, whereas the lines of force on the outside are in opposite directions. This is shown in Fig. 3-15. Due to the basic law that lines of force of like direction repel each other, a force of repulsion exists between the facing lines, and this is transmitted to the conductors as a whole and tends to move them apart. On the other hand, the magnetic loops which advance in opposite directions will produce resultant fields which will be weaker on the outside than on the inside. Consequently, the repelling force issuing from the field between the wires will experience no opposition in moving the two wires apart.

The purpose of showing the action between the fields surrounding two current-carrying conductors is useful, although neither the focusing nor the deflection phenomenon in an electromagnetic type of cathode-ray tube involves the motion of a current-carrying conductor. The reason behind this presentation is to show the manner in which two magnetic fields may combine to produce a strengthened part and a weakened part of the field. And, building upon this, to show how, when such a situation prevails, physical motion can be developed. In this connection it is important to bear in mind that motion on the part of the *conductors is the result of forces applied to the moving charges in these conductors*. If you could envision the currents in the two wires of Figs. 3-14 and 3-15 as being streams of electrons which did not have metallic conducting paths, their physical behavior would be the same as that of the wires. The motion of the wire as a whole is nothing more than a physical manifestation of the magnetic forces being applied to the moving charge.

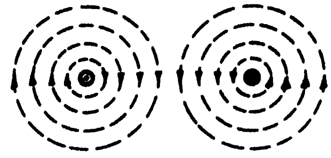


FIG. 3-15. — Magnetic fields around conductors having current flow in opposite directions. Strengthened field between conductors tends to force them apart.

Moreover, as far as the combination of magnetic fields is concerned, the general sense of what has been shown is applicable, if the individual fields are of different character — that is, if one is uniform and the other is nonuniform. For example, Fig. 3-16 shows a uniform field, the origin of which is not of immediate concern, but which has a direction as indicated. Located in this field is a current-carrying wire, which is so oriented as to be at right angles to the lines of force. Thus two fields are active in the area, the fixed uniform field and the circular field due to the moving charges in the wire. The individual fields, as each would be if the other were not present, are shown in Fig. 3-16A, and the resultant field is shown in Fig 3-16B.

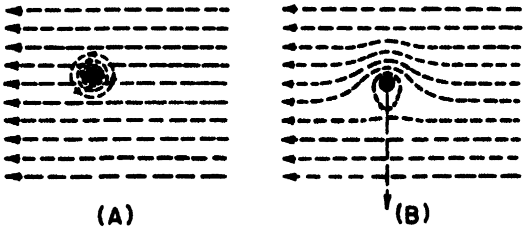


FIG. 3-16.—Current-carrying conductor in uniform magnetic field. Individual fields appear in (A), while resultant field and motion are shown in (B).

Examining the individual fields (Fig. 3-16A), we note that the direction of the lines of force in the uniform field and those around the wire, due to the current in the wire, are the same above the wire and opposite below the wire. The field above the wire is thereby strengthened and the field below the wire is weakened, as shown in Fig. 3-16B. In accordance with the laws previously stated, the wire as a whole will feel a force tending to push it downward, as indicated by the arrow.

If the current through the wire is reversed in direction, the relative positions of the strengthened and weakened parts of the combined field will be reversed, and the wire as a whole will feel a force tending to push it upward. *In either case the direction of the force, and therefore the motion, experienced by the conductor in the combined field is perpendicular to both the motion of the charges and the direction of the field.*

As you will see later, this action is the *basis of electromagnetic deflection* of the electron beam in the cathode-ray tube. The action described above should not be too difficult to recognize as being deflection; all we need do is to visualize the conductor as a stream of moving charges without benefit of a wire as the confining boundary.

Motion of the Electron in a Uniform Magnetic Field

It is conceivable that an electron may be propelled into a uniform field from various directions. It may enter the field at right angles to the lines of force, parallel to the lines of force, or at some intermediate angle between 0° and 90° with respect to the lines of force. We have already gathered some impression of what happens when the direction of the moving charge (in a wire) is at right angles to the line of force.

Let us replace the conductor with an electron which is propelled into the field at right angles to the lines of force, and along an axis X-Y. This is shown in Fig. 3-17 as a view in which we are looking *along* the lines of force into the page. We will assume that whatever projects the electron into the field is active in causing it to traverse the field, which is comparatively short. We know that an attribute of the motion of the charge is a magnetic field which surrounds the moving charge, and if we assume the field to have the direction shown, the resultant field around the moving charge will be stronger *above* the moving charge and weaker *below* it. Therefore, a constant force tending to push the charge downward always will be present, as long as the charge continues to move in the same direction within the field.

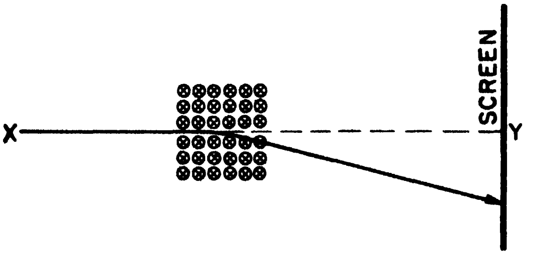


FIG. 3-17.—Deflection of electron projected into short, uniform, magnetic field.

It is important at this time to describe these forces in a general manner because of later needs. When we say that the charge will feel a force downward, it is an expression of a specific direction under the specific conditions stated. A more general statement would be that the moving charge feels a force which is at right angles to its velocity and to the direction of the field. Because of the directions involved in this case, the force is downward. If the direction of the uniform field were the reverse of that shown, the resultant motion would be upward. In either case, the meaning of downward or upward is with respect to the axis; since the electron is advancing forward through the field, the change from its path, the axis X-Y along which it entered the field, is the equivalent of *deflection* of the moving charge.

Because the field is axially short and the electron velocity is high, the electron leaves the field and its influence shortly after entering it, and, when it does so, it travels along a new path, shown in Fig. 3-17. While in the field, the electron describes a curved path, but when it leaves the field, its path is a straight line toward the screen which it would eventually strike. The point on the screen which this electron would strike with respect to the axis X-Y is a function of the angle of deflection at the outer edge of the field and the distance between the center of the field and the screen. Obviously, for any particular angle of deflection, the greater the distance to the screen, the greater is the amount of displacement of the point where the electron strikes relative to the center of the screen. This, naturally, applies equally to deflection in any direction.

If for the moment we assume that, instead of a single charge, a stream of electrons is shot into the field and the beam is stopped at a screen, the system as a whole corresponds in general to what takes place during electromagnetic deflection in a cathode-ray tube. What we have illustrated is deflection in a limited direction — that is, upward and downward relative to the axis, but this limitation is purely a matter of system construction. By suitable design of the system, simultaneous deflection vertically and horizontally can be attained, with such resultant motion of the electron beam as is determined by the relative intensities of the two fields which are acting in quadrature (at right angles to each other).

This is illustrated in Figs. 3-18A, B, and C. In the first part of the illustration, an electromagnetic deflecting assembly is shown, mounted around the neck

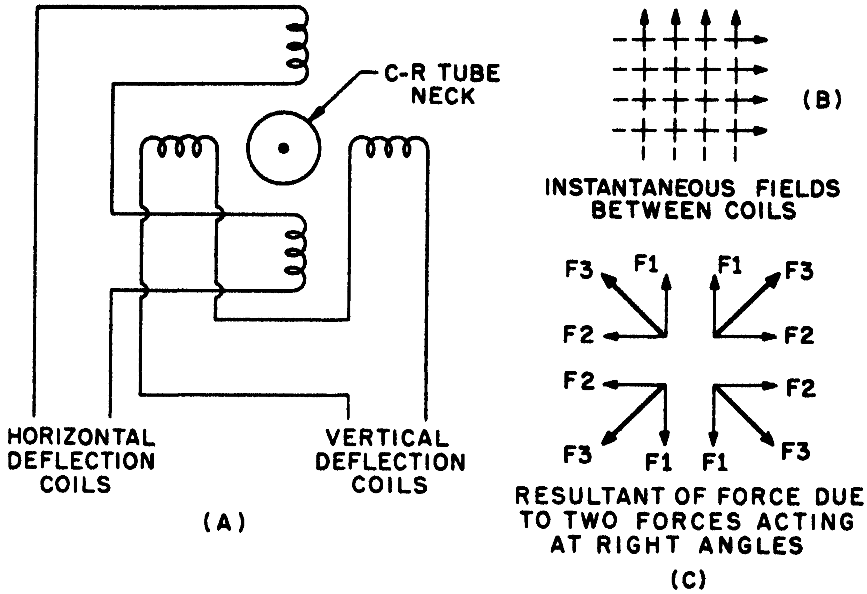

C-R TUBE NECK
HORIZONTAL DEFLECTION COILS
VERTICAL DEFLECTION COILS
(A)
(B)
INSTANTANEOUS FIELDS BETWEEN COILS
F3
F1
F1
F3
F2
F2
F2
F2
F3
F1
F1
F3
RESULTANT OF FORCE DUE TO TWO FORCES ACTING AT RIGHT ANGLES
(C)

FIG. 3-18.—Position of deflection coils relative to cathode-ray tube (A), individual fields of coils at any one instant (B), and four possible resultant field directions due to equal fields acting at right angles (C).

of the tube. The assembly consists of two pairs of windings, one pair for vertical deflection, and one pair for horizontal. The members of each pair are connected in series, and a magnetic field, proportional to the current in the windings, exists between the facing ends of each pair of coils. The axis of each pair of windings intersects the axis along which the electron beam, shown by the heavy dot, advances through the tube. Thus, two magnetic fields occupy the space between the coils, and are at right angles to each other, as illustrated in Fig. 3-18B.

It is understood that lines of force cannot cross each other, so when we say that two fields exist in the space, we really mean that there is a single field which is the resultant of the two component fields. The direction of action of the resultant field is determined by the relative intensities of the two fields, which in turn means the relative intensities of the currents flowing through these windings. When the electron stream passes through the field between the two pairs of windings, it is deflected in a manner determined by the simultaneous action of forces acting at right angles to each other, the four cases of which are shown in Fig. 3-18C.

The magnitude and direction of the resultant force, F_3, which will deflect the electrons is, as has been stated earlier, a function of the relative intensities of the field strengths, or magnetic forces, of F_1 and F_2; in this application, the fields are always at right angles to each other. The coils which are oriented perpendicularly to the earth's surface produce a deflection at right angles to that orientation, or in the horizontal direction; the coils which are positioned parallel to the earth's surface produce a deflection in the vertical direction.

By itself, each deflecting system would produce a deflection such as that shown in Fig. 3-17, which, for one, would be either upward or downward, and, for the other, either to the left or to the right. Depending upon the manner of application of the tube, the deflecting system is used in quadrature, or in either of the two separate directions. Examples of both will be shown later.

Factors Which Determine the Path of the Electron in a Uniform Field

For the individual who may be interested in the mathematics of the motion of the electron in the field shown in Fig. 3-17, the following is offered.

An electron, having mass m and charge e, enters a uniform magnetic field H at a velocity v, and at an angle θ with the field. The force on the electron is

$$f = eH(v \sin\theta) \qquad (3\text{-}3)$$

If the radius of the curved path described by the electron is r, then the equation of motion is

$$\frac{m(v \sin\theta)^2}{r} = eH(v \sin\theta) \qquad (3\text{-}4)$$

or the centrifugal force on the electron equals the force produced by the field, and the radius

$$r = \frac{m(v \sin\theta)^2}{eH(v \sin\theta)} = \frac{m(v \sin\theta)}{eH}. \qquad (3\text{-}5)$$

Since $\sin 90° = 1$

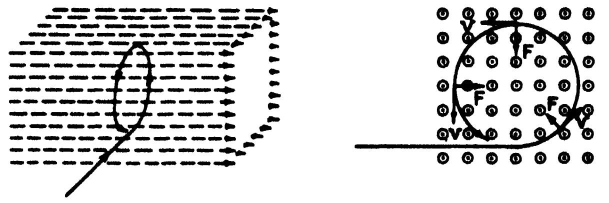
$$r = \frac{mv}{eH} \qquad (3\text{-}6)$$

The quantities m, v, e, and H are constant, therefore the radius r is constant.

If we interpret these factors into magnitude of deflection, then it is evident that: (1) the lower the velocity of the electron per unit field strength, the smaller the radius and hence the greater will be the amount of deflection; (2) the greater the intensity of the field for any one velocity of the electron, the smaller will be the radius, hence the greater will be the deflection; and (3) the longer the time spent within the field (which condition is related inversely to the velocity of the electron and in proportion to the axial length of the field), the greater will be the deflection.

Circular Motion of the Electron

Let us assume that the electron is projected into a long field at right angles to the lines of force, instead of into a short-axial magnetic field. From what was said before, we know that the electron will experience a constant force which is acting perpendicularly to the direction of motion. Since the electron remains in the field, and at every point feels this constant force, the path becomes a circle, as shown in the two views of Figs. 3-19A and B.



(A) FRONT VIEW (B) SIDE VIEW

Fig. 3-19.—Circular motion of electron projected perpendicularly into uniform magnetic field.

The path of the electron, while describing the circle in Fig. 3-19A, is initially upward from the axis along which the charge entered the field. This is due to the relative directions of the field and moving charge. If either were reversed, the electron path would be in the opposite direction. In either case, the radius of the circle is determined by the same factors which controlled the radius of the curved path followed by the deflected electron in Fig. 3-17.

Helical Motion of the Electron

So far we have shown but one angular relationship between an electron projected into a uniform field and the field, that for 90°. Obviously, an electron may be projected into a field along a path which is parallel to the lines of the force — that is, at an angle of 0° to the axis, or at any angle less than 90°. In the first case, the charge will feel no force and will continue to move in a straight line through the field as shown in Fig. 3-20.

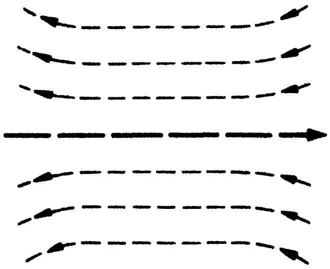


Fig. 3-20.—Path of electron projected into uniform field along lines of force (heavy line) unchanged by field.

However, if the electron enters the field so that it crosses the lines of force at some angle between 0° and 90°, as shown in Fig. 3-21, a composite motion, partly circular and partly linear, will result. Again the field is assumed to be substantially long. Upon entering the field, the electron velocity v_3, can be resolved into two components, a component v_1, which is at right angles to the field, and a component v_2, which is parallel to the field. (See Fig. 3-13.)

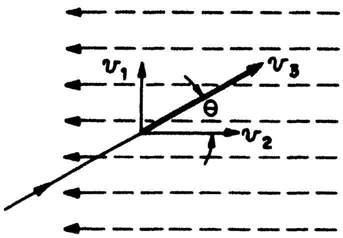


Fig. 3-21.—Vertical and horizontal components of velocity of electron, having velocity v_3, projected into uniform magnetic field at angle θ.

We know that when the charge moves at right angles to the lines of force, a curved path is developed, and when it moves parallel to the lines of force, it will travel unaffected by such motion. So, as the consequence of the component v_1 (Fig. 3-21), the electron will, because the field is exerting a deflecting force, attempt to describe the circular path. At the same time, the component v_2 is causing the electron to advance linearly through the field, so that what would be a circular path for the electron without the v_2 component now becomes a *helical* or *corkscrew-shaped* path as shown in Fig. 3-22. Restating this condition, the combination of circular and linear motion produces the helical path.

As in the case of the electron which described the circle in Fig. 3-19, the direction of rotation of the helix is determined by the relative directions of the advancing electron and the lines of force in the stationary field. If we look in the *direction of the lines of force* (not necessarily in the forward direction of electron motion) the circular motion of the electron will be *clockwise*. From this, it follows that if the forward

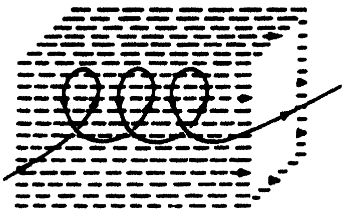


FIG. 3-22.—Helical motion of electron, having both vertical and horizontal velocity components, in uniform magnetic field.

motion of the electron is in the direction of the magnetic field, the path will be an ordinary right-hand corkscrew.

The radius of the turns is proportional to the velocity of the electron cutting across the field. For an electron coming absolutely straight along a line parallel to the direction of the field, there is no cross-wise component; this means that the turn-radius is zero. In other words, there are no loops, and the motion continues in a straight line. For other electrons, not quite along the field direction, there will be some spiraling, and the greater the angle, the greater the loop size; this is due to the fact that a greater angle between field and electron path gives a greater transverse velocity component.

Electromagnetic Focusing of the Electron Beam

The action portrayed in Fig. 3-22 is the essence of electromagnetic focusing of emitted electrons into a concentrated beam. In the practical case, the stationary electromagnetic field may not be exactly as has been shown. The field may not be the long field we referred to; it may be short, axially, yet a similar corkscrew path will occur, and electrons issuing from a point and entering the field at different angles will be brought to a single point, or focus.

In Fig. 3-23 is shown a short magnetic field which might exist in the core of a hollow focusing coil. The field within this space is uniform, although it need not be so to produce focusing action. Not all focusing coil structures are of the type shown; some coils are surrounded by an electromagnetic shield which has a very narrow slit on the core side. The magnetic structure is unimportant, however, since fundamental principles are being discussed. The axis of the system is *X-Y*, and the point *X* also is assumed to be the origin of three electrons *a*, *b*, and *c*. At the extreme right is an imaginary screen which the electrons strike at the end of their path.

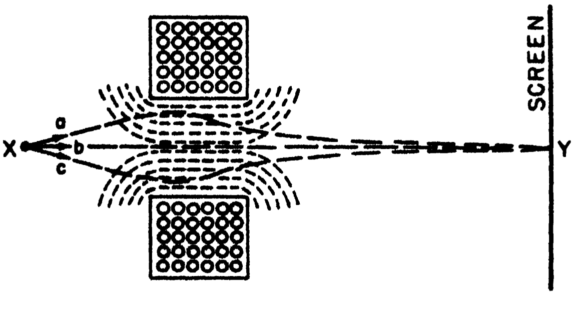


FIG. 3-23.—Focusing of electron beam by uniform magnetic field within core of focus coil.

One of these electrons *b* advances along the axis, and travels to point *Y* along a straight line, the axis of the system. Electron *a* penetrates the field at an angle to the lines of force; after being acted upon by the magnetic force, it emerges from the field and advances in a straight line toward the screen, where it meets electron *b*. In like manner electron *c*, which enters the field at the same angle as *a* but on the opposite side of the axis also completes a spiral path like that of electron *a*, and emerges from the field, continuing in a straight line toward the screen, where it meets electron *a* and *b* on the axis *Y*. Obviously, the straight-line path of the electron after it emerges from the field is determined by the spiral developed within the field. In this connection we assume that the limits of the field are distinct and sharply defined.

If we visualize point *X* along the axis as being an electron beam cross-over point, such as we have described in connection with the first focusing lens in the electrostatic focusing system, it stands to reason that many more than three electrons will leave it and advance into the field at different forward directions. Also the angles formed by these initial velocities and the field will vary between 0° and some amount less than 90°. However, that which was shown to be the condition surrounding electrons *a* and *c* in Fig. 3-22 will be found to be true in the case of any number of charges which enter the field at differing angles. Consequently, all the electrons which enter the field at angles greater than zero degrees, will go through spiral paths of such dimensions that they will converge upon a single point along the axis. By proper design of the focusing system, this point can be placed at any desired distance from the termination of the field. In our case it is upon the screen.

If we view point *X* as the cross-over point, then the point at which the electrons in the beam are brought to a focus is the ***image of the cross-over point***. It is important to emphasize this, to dispose of the impression that the spot which appears upon the screen of the tube, where the undeflected beam strikes the surface, is the image of the cathode.

Such an electron lens formed by a magnetic field is a convergent lens and behaves much like a convergent

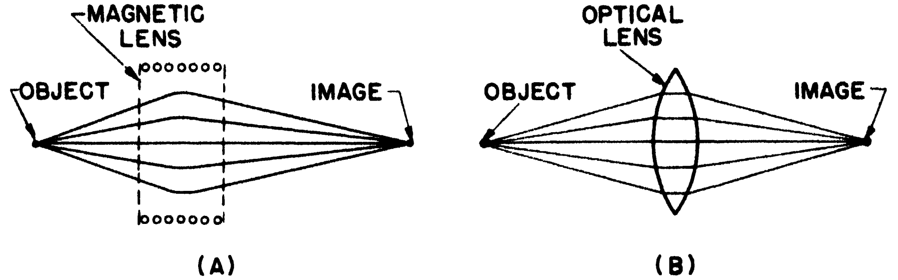


Fig. 3-24.—Similarity between focusing of divergent electrons by electron lens (A) and focusing of divergent light rays by optical lens (B).

lens used in light optics. It gathers divergent electrons and brings them to a focus, just as the convergent light lens gathers divergent rays of light and brings them to a focus. It should not be forgotten that the image point also is a cross-over point, beyond which both electrons and light rays would again pursue diverging paths. This is shown in Figs. 3-24A and B. A number of spiral paths for electrons which entered the field at different angles are shown in Fig. 3-24B; the fact that these paths are curved while under the influence of the field should not be confused with the previous statement to the effect that after the electron leaves the field it advances along a straight-line path. The similarity in focusing between the magnetic lens and the optical lens is still evident.

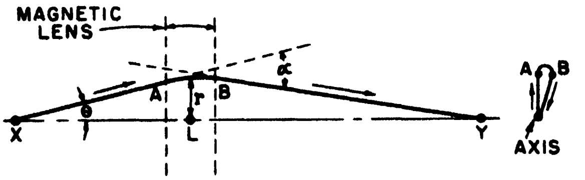


Fig. 3-25.—Path of electron under influence of short magnetic lens. Spiral motion between points *A* and *B* is shown in end view.

Part of the mathematics of focusing may be of interest to some readers. From point *X*, in Fig. 3-25, an electron starts out at an angle θ, and enters the magnetic field at *A*. In passing through the field, it takes the curved path *AB*, and is deflected through the angle α, back toward the axis. As may be seen in the end view, if the curve is much too sharp or not sharp enough, the returning path may fall wide of the axis *X-Y*. But for approximately correct values, small variations in the curvature will change the distance *LY*. Now, for small angles of electron divergence — and this is necessary in both theory and practice — any angle, expressed in radians, is equal to its tangent. (A radian is an angular measurement, equal to 57.3°.)

Therefore

$$\angle AXL = \frac{r}{XL} \tag{3-7}$$

$$\angle BYL = \frac{r}{YL}\ . \tag{3-8}$$

Since, from the relationship between the two triangles

$$\angle AXL + \angle BYL = \angle \alpha \tag{3-9}$$

then

$$\angle \alpha = \frac{r}{XL} + \frac{r}{YL} = r\left(\frac{1}{XL} + \frac{1}{YL}\right). \tag{3-10}$$

By the use of calculus, it is found that α is proportional to r, or

$$f\alpha = r \tag{3-11}$$

where f is the factor relating α to r. Substituting this in equation (3-7) above

$$\alpha = f\alpha\left(\frac{1}{XL} + \frac{1}{YL}\right) \tag{3-12}$$

or

$$\frac{1}{f} = \frac{1}{XL} + \frac{1}{YL}\ . \tag{3-13}$$

The distance *XL* corresponds to the object distance in optics, and *YL* to the image distance, designated d_o and d_i respectively. We may write

$$\frac{1}{f} = \frac{1}{d_o} + \frac{1}{d_i} \tag{3-14}$$

for the magnetic lens, corresponding to the same equation for an optical lens. In this case, f, which is the factor relating deflection angle to this distance of the electron paths from the axis, is dependent upon *H*, the magnetic field strength. In optics, f is the focal length. Effectively, then, we have a magnetic lens for electrons whose focal length can be varied by changing the magnetic field strength, within certain practical limits.

Since all the electrons leaving *X* at a small angle to the axis, are brought to the same point, *Y*, the point is a true focus, and the comparison with an optical lens is quite appropriate.

Typical Electromagnetic Focusing System

Referring back to Fig. 3-1, we may now analyze the operation of a typical cathode-ray tube which makes use of electromagnetic focusing and deflection. This is a representation of a simple system, not necessarily a sample of any one particular piece of equipment. The cathode, or source of electrons, is enclosed by the control grid. Next to the control grid, is a screen grid; it is obvious that with a voltage applied to the screen grid, an electrostatic focusing lens is formed between the apertures of the control grid and the screen grid. Here the cross-over point in the partially formed electron beam, which is the "object" that is reproduced as the spot image, can be seen. The baffle plates in the screen tend to prevent entry into the magnetic lens of widely divergent electrons, thus simplifying the focusing.

Near the screen grid is the opening of the accelerating anode. In practice, this element is usually a conductive coating on the inside of the tube. The high voltage used to accelerate the electron through the focusing and deflecting systems and toward the screen of the tube is applied to the anode. The deflection area is located between the focusing system and the screen.

The partial focusing action, which takes place between the control grid and the screen grid, has been described previously. A stream of electrons issues from the cross-over point thus formed, and is projected into the focusing area. The position of the magnetic field, generated by the focusing coil, with respect to the stream of electrons, usually is adjustable so that the succeeding cross-over point may be made to coincide with the screen.

CHAPTER 4

MECHANICAL CHARACTERISTICS

There was a time, when high-vacuum cathode-ray tubes first made their commercial appearance, that few types were available. They were, in the main, small- and medium-size tubes; and while it is true that both electrostatic and electromagnetic deflection types were available, the former was by far the more common. This may seem strange, since the electromagnetic method of deflection is the older of the two; yet the early needs for cathode-ray tubes were most numerous in connection with applications in which electrostatic deflection was found easier to use.

Since that time many changes have taken place. Not only has the scope become much more popular, but so many new uses for the cathode-ray tube developed in connection with the war effort that the range of types has expanded very greatly. Then along came commercial television and still further increased the field of the tube. Finally, the needs of research as well as education have resulted in the creation of multi-electron-gun tubes, which still further increased the number of tube types.

Under the circumstances, it seems quite reasonable that we should present, in full detail, the points of similarity and dissimilarity between these various types. Although it is true that as far as complete cathode-ray scopes are concerned, the number of types remains restricted, the span of this text embraces more than just the conventional radio and television maintenance industry. Therefore all types known to be within the category of high-vacuum cathode-ray tubes, as well as their applications, merit examination.

Comparison of Cathode-Ray-Tube Types

To the eye many different types of cathode-ray tubes look alike. Yet they may differ in so many ways that interchanging one with the other is out of the question. On the other hand, many tubes which appear different may have sufficiently similar characteristics as to permit interchanging with little effort in order to take advantage of the more salient features of one type over the other. As a matter of fact, cathode-ray tubes *cannot* be judged by physical appearance. Certain features may *appear* to be the same, but external appearances invariably are useless as a means of evaluating comparative utility or suitability.

The correct method of comparing cathode-ray tubes is by means of an analysis of the electrical characteristics of the tubes, and this only after determination of the adequacy of the mechanical characteristics. Mechanical characteristics are of paramount importance when the interchanging of tubes in already existing equipment is contemplated. The feasibility of using a different type of tube in an existing piece of equipment is dependent first upon its ability to physically accommodate the new tube.

After this has been decided, then, and only then, is it justifiable to examine the electrical characteristics. Many different items must be analyzed, not only those which might seem the most important, such as heater voltage and current. Tube operating parameters must be critically checked to establish the suitability of the tube for use with circuitry in the unit. Many individuals have experienced great disappointment after making a change in the type of display when they discovered that the available operating voltages did not permit taking full advantages of the tube with the larger screen; or that the amount of amplification available in the associated amplifiers was insufficient to fully utilize the new tube.

The designer or constructor of new cathode-ray-tube equipment is not faced with all of these limitations, since he is building something new; yet he too must make a careful and complete analysis in order to incorporate such design in the related apparatus as will enable the realization of the full advantages of the tube which he contemplates using.

On the surface it may appear that simple circuit modifications will permit substitution of cathode-ray tubes in equipment. Sometimes this is possible, but ordinarily, the reverse is true. Even after the circuit changes have been made and the equipment is functioning, it is found that either the necessary regulation does not prevail, the proximity of the tube to the components present in the system gives rise to undesirable effects, or the tolerances of the components are not strict enough to properly fit the new needs.

It would be foolish to say that one cathode-ray tube cannot be substituted for another, because this is not so. Replacements with improvements are made for that purpose; however, after many years of experience, we can say without fear of contradiction that interchanging of cathode-ray tubes which have like physical or mechanical characteristics but different electrical characteristics presents considerable difficulty. The simplest type of interchangeability exists when all dimensions and characteristics are the same, except for the fluorescent screen material. Such replacement often becomes necessary in photographing certain oscilloscope displays. Although the technique of photography and the range of film speeds and color response have improved to the point where good photographs of most traces may be obtained from virtually any type screen, certain screens, designed specifically for high-speed transients, are generally required for photographic recording of these patterns.

The comparison of cathode-ray tubes is not difficult, since all of the data are available; but it is imperative that all of the facts of the problem be understood. The understanding of mechanical characteristics poses no difficulty; the comprehension of the electrical characteristics demands the greatest attention.

With this in mind, we shall devote this chapter and the next three to a study of the details of cathode-ray tubes. We must, however, comment that the extent of the subject covered will be within boundaries set by the fields of application which we are embracing in this text, and by tubes used in such systems. First of all, only high-vacuum type tubes will be dealt with — the tubes will be found in devices known as cathode-ray oscilloscopes or oscillographs and synchroscopes. Second, we shall include all of the types of tubes which are used for display of information in connection with industrial, medical, and scientific activities, with the exception of the very-high-voltage oscillograph and the electron microscope. Third, we shall include the displays which present information derived from measurement of time or direction of travel, or both, of electromagnetic waves; this classification includes systems of military and commercial importance, particularly navigational aids.

The cathode-ray tubes employed in television receivers will receive full attention, although it is inadvisable to set up a special category for these tubes. In general, they are like the tubes employed in other applications, except perhaps for differences in the screen materials because of the need for certain persistence characteristics.

The Basic Cathode-Ray Tube

For comparison of the many varieties of tubes, it is essential to have, as a starting point, a basic cathode-ray tube. Fortunately this is possible, because no matter what the type or purpose of tube under consideration, whether electrostatic, electromagnetic, or combination, all have certain features in common. For example, the internal construction of all cathode-ray tubes can be divided into four basic operational sections. It is possible to subdivide, and list many more, but these four are the cardinal sections associated with the action upon the electron beam. They are:

1. The source of the electrons which are eventually concentrated into the beam.
2. The assembly of elements whereby the emitted electrons are controlled, focused, and accelerated toward the screen.
3. The assembly of elements whereby the deflecting forces are applied to the beam before it reaches the screen.
4. The screen of the tube.

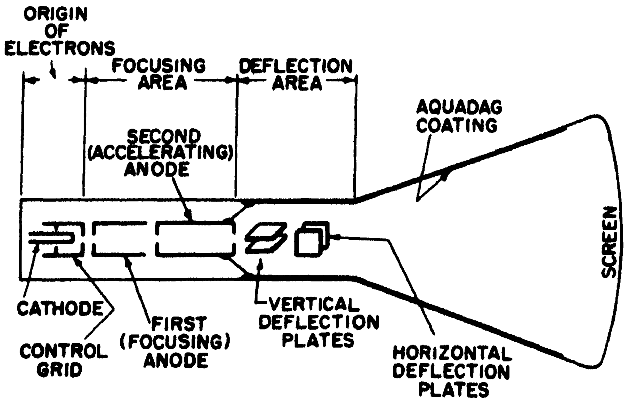


Fig. 4-1.—Organization and function of elements in electron gun of basic electrostatic cathode-ray tube.

If we illustrate the organization of these four sections as they appear in the basic cathode-ray tube, we have the three pictures shown in Figs. 4-1, 4-2, and 4-3. The first of these is the conventional electrostatic cathode-ray tube; the second one is the conventional electromagnetic tube, and the last is a combination of both electrostatic focusing and electromagnetic deflection. Each illustration contains notations identifying the action of the sections.

It is to be remembered that these are simple representations of three general varieties of cathode-ray tubes and are useful only from the viewpoint of the organization of the basic functional section of the device. Later on as the descriptions of the tube types develop, the variations from these basic types will become evident, but even the modified versions will be found to conform with the general pattern indicated by these elemental types.

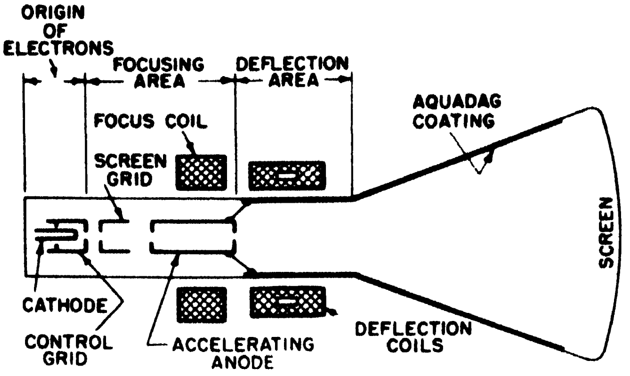


Fig. 4-2.—Organization and function of elements in electron gun, including external focus and deflection coils, of basic electromagnetic cathode-ray tube.

The Electron Gun

At the moment, we are not concerned with the actual construction of the electron gun, which functionally embraces items 1 through 3 of the list given, or with the physical construction of the different sections. What we desire to stress is that the *standard cathode-ray tube,* regardless of type, employs a *single* electron gun which contains the electron emitter and a means for controlling the density of the electron stream. From this point on, the actual arrangement of elements for *concentrating* the electrons into a beam, that is, for focusing, is a matter of the individual type.

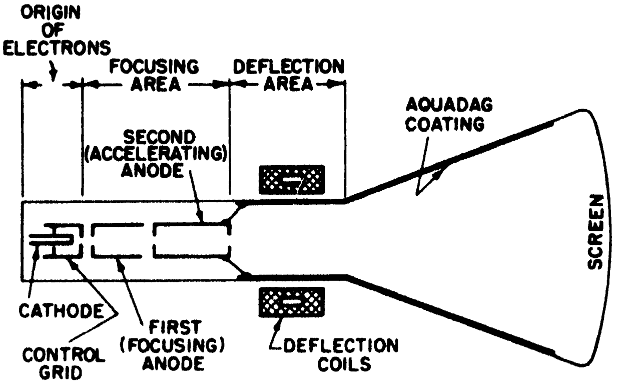


Fig. 4-3.—Organization and function of elements in electron gun of combination electrostatic-electromagnetic cathode-ray tube.

The electrostatic tube employs a gun which includes not only the focusing electrodes, in addition to the emitter and the control electrode, but also the means for accelerating the beam on its way toward the screen, and the elements required for deflecting the beam horizontally and vertically. This is shown in Fig. 4-1. Strictly speaking, the electron gun does not really include the deflection elements, but it has become standard practice to assemble and mount these as a unit because it is most convenient for manufacturing. It seems that it is easiest to mount the deflection elements when assembling the other parts of the gun, especially since the orientation of the control and focusing electrodes is important with respect to the centering of the two pairs of deflection plates. By working with a completely integrated unit, physical alignment of the elements is improved. But ordinarily when we discuss the electron gun used in the electrostatic type of cathode-ray tube, the deflection system is not included.

In the electromagnetic variety of tube shown in Fig. 4-2, wherein both focusing and deflection are accomplished by electromagnetic means, as generally understood the electron gun is limited to the electron emitter, the control element, and a part of the beam accelerating system. The focusing coil and the deflection coils are apart from the gun; in fact they are located outside the tube, with the deflection coils positioned at a point definitely removed from gun location. The means provided for individual positioning of the focusing and the deflection coils so as to attain the best operating conditions are secondary to the organization of the systems within the tube. As was mentioned in the previous chapter, the location of the focusing coil is usually in the proximity of a part of the accelerating element; the full dimensions of this electrode are greater than indicated in the illustration.

In the combination electrostatic and electromagnetic tube, as shown in Fig. 4-3, the electron gun contains the electron emitter, the control element, and the focusing system, as well as an accelerating anode such as is used in the fully electrostatic-type tube. The deflection system is, however, electromagnetic.

Deflection Systems

The deflection system in electrostatic tubes uses two pairs of plates, positioned at right angles to each other, one pair being intended for horizontal deflection and the other pair for vertical deflection. In the electromagnetic tube as well as the combination tube, shown in Figs. 4-2 and 4-3 respectively, the deflection system consists of two pairs of coils so arranged as to produce two magnetic fields at right angles to each other, one pair affording deflection in the horizontal direction, whereas the other pair causes vertical deflection.

The Screen

As to the screen, very little need be said in this preliminary discussion. Each tube has a single screen;

this statement has no implications regarding the make-up of the screen, that is, the kind of material used or the number of layers, but rather that an ordinary cathode-ray tube does not have separate screens.

Accelerating Anodes

Although the accelerating anode is a basic element of the cathode-ray electron gun, it is general practice in modern tube design to modify this somewhat. In most tubes manufactured, the second anode extends beyond the gun region into the cone of the envelope. This is done by coating the inside of the glass with colloidal graphite (aquadag); in many instances, this conductive aquadag coating forms the anode by itself, without any actual metallic element being associated with the electron gun. In the majority of large tubes, the connecting terminal is located near the flare of the cone, if so permitted by the envelope construction. However, in some instances, particularly in small tubes, connection is made through a pin on the tube base.

A different method of electron acceleration makes use of the *intensifier* electrode or *post-deflection* accelerating anode. Its omission from the basic cathode-ray tube of present discussion is deliberate, although its use is becoming more common. Because many tubes do not have this element, it seems better to present it as a modification of the standard construction, rather than as a component element of the basic type of tube. The purpose of the element, or elements as the case may be, is to accelerate the electron beam *after* it has been deflected, causing the beam electrons to strike the screen with higher velocity, and thereby producing a more brilliant trace on the screen. By applying the accelerating voltage after deflection, the effect of increased electron velocity upon deflection sensitivity is kept to a minimum. A more complete discussion will appear later.

Outline of Action Within Basic Tube

The electrons are emitted by the cathode around which is mounted the control-grid electrode. This is a sleeve with an aperture facing the emitting surface of the cathode. The general function of the control grid is to control the density of the electron stream prior to its entry into the focusing area. This it can do sufficiently to cut off the electron flow completely.

Once in the focusing area, the divergent electrons in the stream are concentrated into a beam by whatever means are employed for focusing. All the while, the electrons in the beam are being accelerated toward the tube screen, by the voltage applied to the second anode. While advancing toward the screen, the beam is caused to traverse the deflection area, where the vertical and horizontal deflection forces are applied electrostatically or magnetically, as the case may be. After deflection, the electron beam continues its travel until it strikes the screen, upon which it produces a visible trace.

Such a description of the actions found within the tubes is, of course, extremely sketchy; later on we shall analyze the operation of each of the elements in detail and examine the functional relationship between them, as well as the manner in which each affects the ultimate result.

Types of Cathode-Ray Tubes

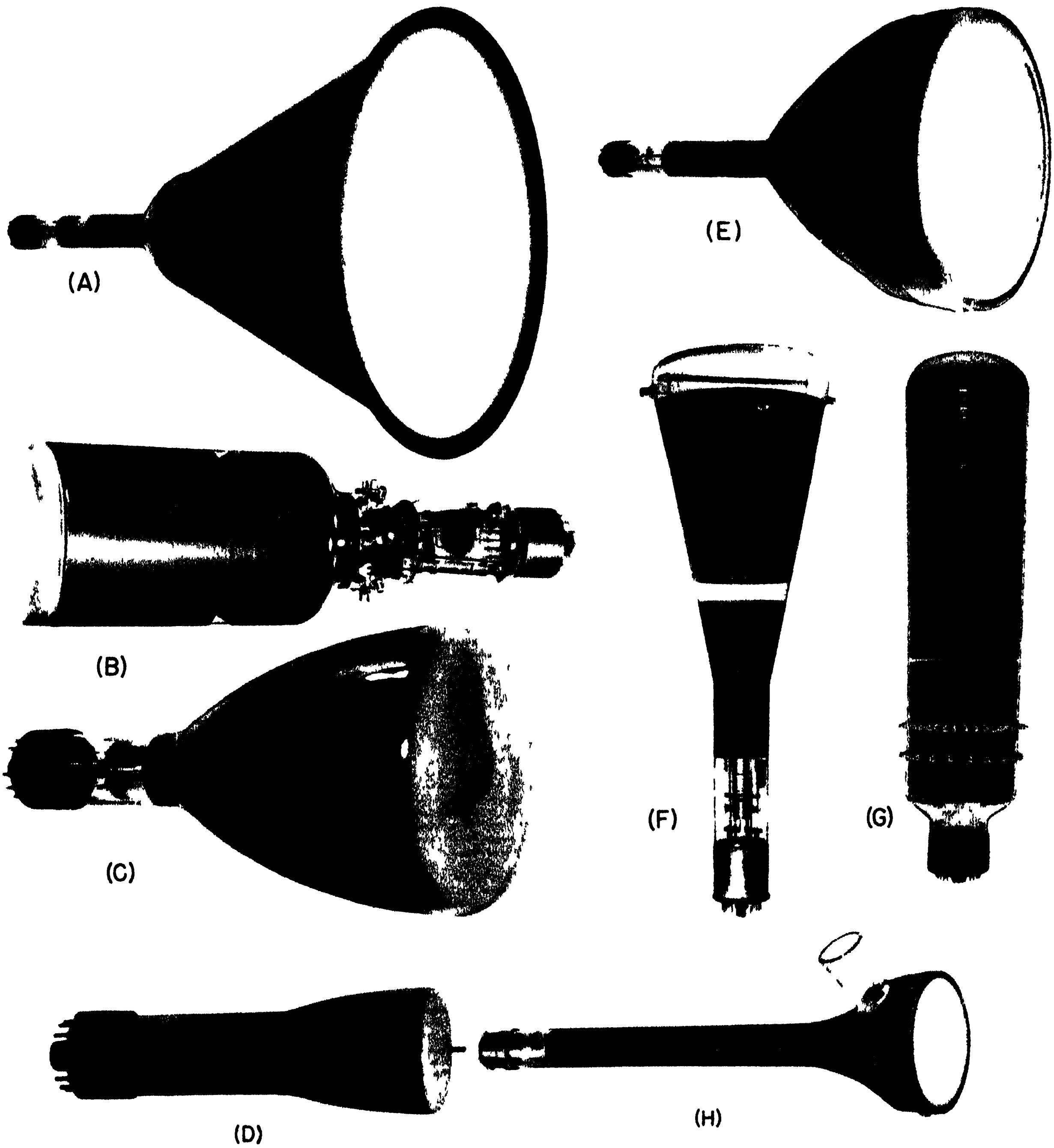
A tabulation of cathode-ray-tube types is based on many different considerations. Not only do tubes look different, but even some which look alike may be of different mechanical construction and possess different mechanical characteristics. Electrically, many almost identical-appearing tubes may have such widely different characteristics as to be totally different tubes. Finally, on the basis of the nature of the screen luminescence, the tubes may differ so much as to make certain tubes totally unsuitable for specific applications. So we find that three distinct classifications may be used to distinguish tube types. In fact there could even be said to be five classifications, if we include the manner of deflection and number of electron guns as further ground for distinguishing tube types.

Type Numbers

The conventional cathode-ray tube manufactured today is identified by an assigned RMA type number which is partially descriptive of the tube. Examples of such designations are the 2AP1, 5BP1, 10BP4, etc.

In these standard type numbers, the *first digit* identifies the *nominal diameter* of the screen in inches. The *first letter* merely identifies the order in which tubes of the same diameter were registered with the Radio Manufacturers' Association. The last *letter-number combination,* that is, the letter "P" followed by a number, identifies the kind of phosphor used for the screen of the tube, thus indicating the color and persistency characteristics.

Consequently, the designation 5BP1 reveals that the tube is five inches in diameter, that it was the second five-inch tube registered with the RMA, and that the screen phosphor is green, with medium-short persistence. Each of these tube properties, especially those concerned with the screen, will be treated in detail subsequently.



Courtesy manufacturers listed

Fig. 4-4.—A representative group of cathode-ray tubes, both general and special purpose They are as follows: (A) National Union 16AP4 metal tube, (B) Du Mont 5SP2 dual-gun tube, (C) National Union 7JP4, (D) RCA 3DP1 J-scope tube, (E) National Union 10BP4, (F) Electronic Tube Corp. 51LN potentiometer tube, (G) Electronic Tube Corp. 7Z10P2 10-gun tube, and (H) Norelco 3NP4 projection tube.

In 1937 a modification was made in the design of the electron gun used in certain tubes and the letter "A" was appended to the type number to show that the tube contained the new gun. Thus, tubes with the same basic type numbers, as for example 5BP1 and 5BP1-A, are alike except for the difference in the gun. The screen dimensions and screen phosphor are the same; for all practical purposes, the two types are interchangeable.

This system of identification is usually applied only to conventional cathode-ray tubes, and not to such special types as the monoscope, image orthicon, some multigun tubes, etc. These are still generally assigned type numbers which are nondescriptive. Prior to the adoption of the number-letter system, conventional cathode-ray tubes were also assigned such numbers, usually in the 900 or 1800 series. However, the need for standardization in nomenclature led to the present system, which as previously stated, offers a partial description rather than only identification of tube type.

Physical Shapes

The most obvious consideration is the shape of the tube or the envelope. This is the outside appearance; the internal arrangement is hidden because of the manner of tube construction, especially the opaque coating on the inside of the bulb. In Fig. 4-4, there are shown various representative types of tubes. Each of the tube types is identified by the manufacturers' type number. A full representation of all brands is not necessary since, due to present-day standardization, tubes bearing like type numbers have like construction and characteristics regardless of make. What we have tried to do here is to show the greatest variety of envelope shapes. The complete specifications — mechanical, electrical, and optical — for every known American-made cathode-ray tube are given in Appendix I.

Fig. 4-4 also shows some mechanical characteristics other than shape. These are discussed later in this chapter. In the meantime, let it be said that there is no *basic* shape, at least not today. There was a time, perhaps fifteen years ago, when all cathode-ray tubes were pear- or funnel-shaped—that is, having a narrow neck, gradually flaring out, and a substantially flat screen surface. Perhaps this construction is most popular even today, but there are so many other types, that we can scarcely consider one shape as "basic".

A unique departure in cathode-ray-tube construction, at least as far as the envelope is concerned, is the so-called metal-type 16AP4, which is shown in Fig. 4-4. Although classed as a metal tube, the construction is a combination of a glass neck, a metal truncated cone, and a glass face for the screen. A cross-sectional view of the envelope is shown in Fig. 4-5A. Being a new development in cathode-ray-tube construction, the device warrants more than usual comment.[1] It is, perhaps, the forerunner for envelope design of the future for all cathode-ray tubes, although at the present time, the standard glass envelope is considered more economical for tube sizes up to 10-inch screens.

Three reasons are advanced for its development: (1) lighter weight, (2) greater manufacturing flexibility, and (3) better optical quality. There are other reasons as well, which extend right to the television set purchaser's pocketbook, but those we have mentioned are sufficient. Manufacturing problems were numerous; foremost among them was the development of the proper seal between metal and glass when the dimensions were as great as the periphery of a 16-inch diameter screen. Glass-to-metal seals have been used for a long time in high-power transmitting tubes, but the diameters were not as great. The selection of the metal was a problem in that certain specific requirements had to be met. The metal chosen, after a great deal of experimentation, as having the proper coeffi-

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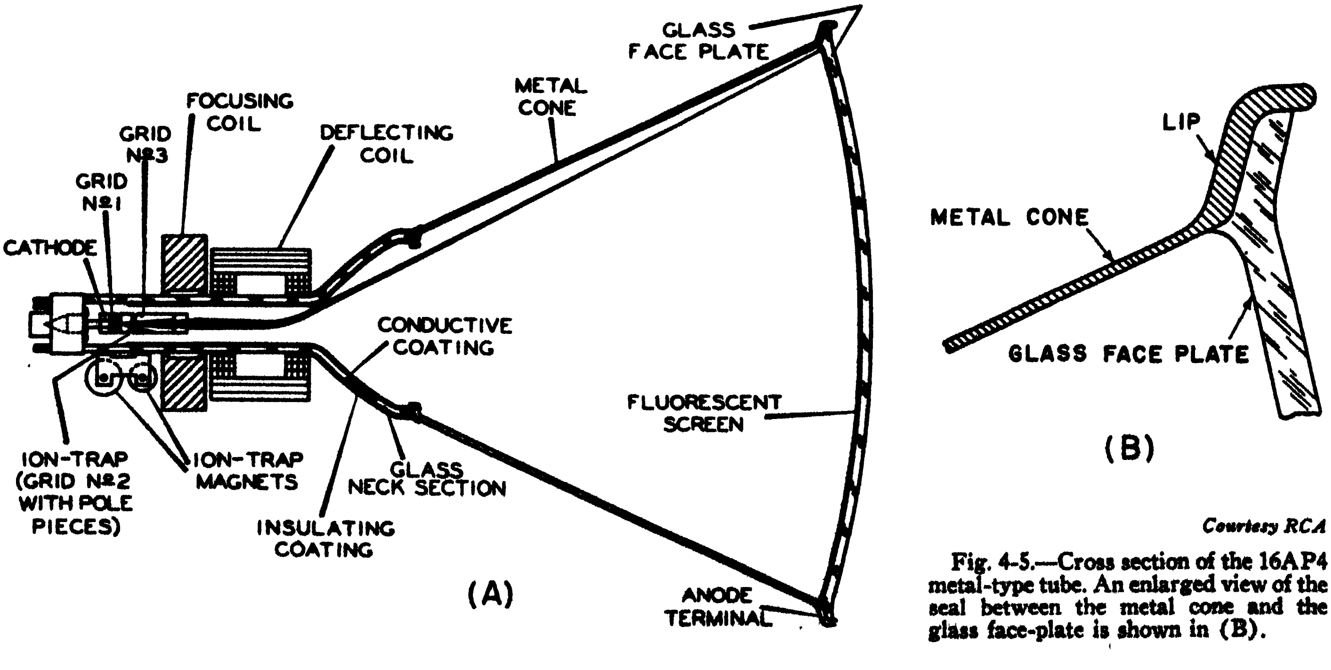
[1] Steir, H. P., Kelar, J., *et al*, "Development of a Large Metal Kinescope for Television," *RCA Rev.*, pp. 43-58, vol. 10, March 1949.



Courtesy RCA

Fig. 4-5.—Cross section of the 16AP4 metal-type tube. An enlarged view of the seal between the metal cone and the glass face-plate is shown in (B).

cient of expansion relative to the glass, allowing a vacuum-tight seal, having high tensile strength, and possessing good corrosion resistance, was an alloy of chrome-iron, known as SAE Type 446.

An interesting sidelight on the construction is the conditions associated with the glass face plate. Normally, the pressure of the air on a surface 16 inches in diameter is slightly under 1.5 tons. For safety purposes, the tube was designed to withstand a pressure of about 4.5 tons. In an all-glass bulb, the loading created by the air pressure is supported by a relatively heavy wall near the maximum diameter of the bulb. In the case of the metal cone, this support is achieved by building up the face-plate-sealing surface of the cone in the shape of a truncated rim-cone supported at one end by a cylindrical rim and at the other end by the main cone. Fig. 4-5B affords an enlarged view of the face-sealing area. Other details of this tube are discussed later under the appropriate subject headings.

Over-all Length

As a general rule, the over-all length of a tube in inches is related to the diameter of the screen. Not that an exact ratio between the two exists, but rather that the longer the tube, the larger the screen and vice-versa. There are, of course, exceptions to this relationship, but generally it is true. For what it may mean as information, among the numerous electrostatic tubes, the approximate ratio of over-all length to screen diameter varies between 2.5 to 1 in the medium length tubes, to about 4 to 1 in the smaller tubes. The higher ratio may be said to apply to tubes with screen diameters up to about 5 inches. As the screen diameter exceeds 10 inches, the over-all length to screen diameter decreases, until the 20-inch tube shows a ratio of about 1.3 to 1. The over-all length of the electrostatic group of tubes ranges from shortest, 4.75 inches, to the longest, 28.5 inches. These figures include the tube pins in the base.

In the case of the electromagnetic tubes, the length-to-diameter ratio is somewhat less, being about 1.4 to 1 in the case of the largest tube, with the 20-inch screen, to about 2 to 1 in the smaller tubes. The shortest tube available is 10 inches long and the longest is about 29.5 inches.

Inasmuch as special needs originate special construction and conformation, these figures should not be taken as absolute limits, for it is possible that by the time the ink in this book dries thoroughly, new tube dimensions may be in use. In fact some multigun tubes exceed the over-all lengths stated above. Because of the difference in over-all length of tubes, if that physical dimension is important, each tube type should be treated separately.

Screen Dimensions

The dimensions of cathode-ray-tube screens are important from the standpoint of display size and readability in oscillographs and similar equipment, and from the standpoint of over-all picture size in television receivers.

In the electrostatic type of tube, the smallest rated screen diameter is 1.3 inches and the widest is about 20 inches. Among the electromagnetic tubes, the smallest screen diameter is 2.5 inches and the largest is 20 inches. This upper limit should not be confused with projected image dimensions in television receiving systems. Picture areas in excess of that afforded by a 20-inch screen are in use, but the tubes on which the image appears usually have screens 5 inches in diameter or less, while the image is enlarged by optical projection.

It is interesting to mention that a casual interpretation of the diameters stated may give one the impression that these screen sizes range from the smallest to the largest in small increments. Such is not the case. The range of diameters for electrostatic tubes can be said to be made up of about eight different fixed dimensions, these being 1, 2, 3, 5, 7, 10, 12, and 20 inches. The same restriction in sizes also exists for the electromagnetic type of tube, in that the full range is composed of tubes of 2, 3, 5, 7, 9, 10, 12, 15, 16, 19, and 20 inches. Obviously, the latter type of tube has a greater number of individual sizes, brought about no doubt by the combined needs of the recent war effort and present-day television.

Identification of screen diameters is accomplished by the tube type number, if it is of the combination digit-letter variety. As explained previously, the first number in such type designation is a statement of the rated diameter of the tube screen expressed in inches. This dimension is the nominal over-all diameter of the tube face, and does not take into account any reduction in the useful area caused by curvature of the screen at its periphery.

In general, tube screens are not perfectly flat, but they are sufficiently so over a substantial portion of their area, as to be the equivalent of a plane surface. However, picture tubes intended for television receiving systems, although not necessarily limited to such use, have flatter screens than the conventional run of cathode-ray tubes. It is equally important to understand that all tubes of like screen diameter do not necessarily afford equally flat surfaces, although tubes of like type number produced by different manufacturers will, in general, afford similar physical features. This is the result of standardization and compliance with joint Army-Navy (JAN) Specifications.

Rated Screen Diameter versus Useful Diameter

Because of the curvature of the face of the tube, especially at the edges, the entire screen does not offer equal utility for display purposes. Neglecting any controls associated with the tube deflection system, it can be said that maximum *useful* screen diameters approximate 80 to 90 per cent of the rated screen diameter. Since the faces of all tubes do not have exactly the same degree of flatness, and since the amount of tolerable distortion is indeterminate, being dependent upon the user, nothing other than a rough approximation of the useful diameter is possible.

The purpose or nature of the display has great bearing on whether the rated screen diameter will be used or if the dimensions of the trace will be limited to about 90 per cent of the screen diameter. The author has found this last figure to be the practical maximum when the trace on the screen is being studied. Thus, if a tube screen has a rated diameter of 5 inches, a good limit for the pattern dimension is 4.5 inches; for a 10-inch screen this would be about 9 inches; for a 3-inch screen it would be about 2.7 inches, and so on.

Naturally it is permissible to stretch this portion, if the distortion which develops around the edges is acceptable, and if the design of the deflection system allows the use of almost the entire screen. In some cases, especially when more than one electron gun is used, the constructional limitations tend to preclude the use of the full diameter of the screen for presentation of information.

The more severe the demand made upon the presentation for study purposes, the greater the care which must be exercised in the dimension of the trace. If distortion due to the curvature of the tube is not recognized, the analysis of the trace can be greatly in error. It has been our experience in oscillographic work, that by and large, trace dimensions should not exceed 90 per cent of the screen diameter. Of course, if only a segment or section of a pattern is to be studied, then it is possible to expand the trace to fill the screen, or even extend beyond the limits, provided that the segment being studied is kept within the 90 per cent limit. This figure allows a safe margin.

In the presentation of television pictures, the RMA standard transmitting aspect ratio of 4 to 3, the ratio of the horizontal to the vertical dimension, is not always reproduced precisely at the receiver. If the picture size were such that no part would extend beyond 90 per cent of the diameter, only about 60 per cent of the screen area would be utilized. In practice, the picture is usually enlarged so that the horizontal dimension is within distortion limits, and the corners run to the edge of the tube, or are lost altogether; in such a case about 70 per cent of the tube area can be used. Further gain in picture size can be obtained only at the expense of additional distortion and loss in portions of the transmitted picture.

Tube Basing

The basing of cathode-ray tubes has been standardized only with respect to identical tube types made by different manufacturers. All tubes which bear identical type numbers have the same type of basing. However, not all tubes of like screen diameter or like over-all length are of the same type nor are they necessarily based alike. Therefore, sockets for cathode-ray tubes must be properly selected. In this connection, it is necessary to introduce a subject somewhat prematurely, but only to a limited extent. This is the matter of the number of electron guns within the tube envelope.

In the usual run of single-gun tubes, which can be considered to be the standard type, ten different types of tube bases are employed. These are as follows:

Medium 5-pin
Medium 7-pin
Octal 8-pin
Magnal 11-pin
Duodecal 5-pin
Duodecal 7-pin
Duodecal 11-pin
Duodecal 12-pin
Diheptal 12-pin
Peripheral 12-contact.

Special purpose tubes such as the image orthicon, the RCA monoscope and others make use of two additional bases, the Medium 6-pin and the Diheptal 14-pin.

If we examine cathode-ray tubes which have more than a single gun, still other tube bases are used. For example a two-gun electrostatic cathode-ray tube with 5-inch screen made by one manufacturer uses a 12-pin base, whereas another two-gun tube with 12-inch screen makes use of a 20-pin tube base. Still another multigun tube uses a 40-pin base. Identification of the specific type of basing for each tube is given in Appendix II.

The differences in tube basing demand proper correlation between the base and the socket. Tubes which have like screen dimensions and over-all length, and even similar characteristics, which to all intents and purposes are interchangeable, may be found to use substantially different arrangements of electrode connections. That is to say, the same element in each tube does not always connect to the same numbered pin on the base. This calls for proper correspondence between the elements within the tube and the circuitry connected to the socket. Obviously, this is no problem when handling complete equipments which already contain the appropriate cathode-ray tubes, but does require attention, when, for some reason, one type of tube is replaced by another.

Apart from the number of contacts and the differences in diameter, the general conformation of the tube bases is the same except for the peripheral variety. In this type, shown in Fig. 4-6, the contacts are located around the periphery of the base, hence its name. The socket used with such a base also is shown in Fig. 4-6 and as may be evident from the illustration, the connections are located around the inside wall of a flange of the dielectric material.

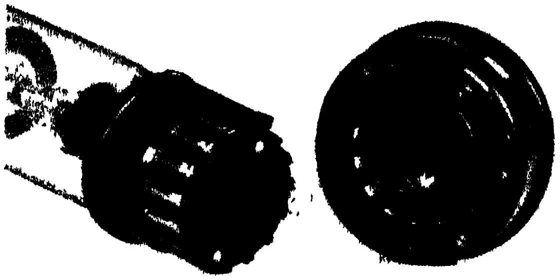
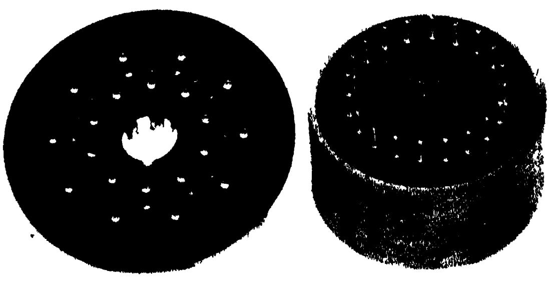


Fig. 4-6.—Peripheral 12-contact tube base and socket.

A tube base with 20 pins and the special type of socket which it requires are shown in Fig. 4-7. The base contacts are assembled to form a circle, despite the relatively great number of pins. The socket is of interest in that, while the points of contact with the tube base pins form a circle, the connecting clips are arranged in two rows, an outer row and an inner row; in that way maximum separation and insulation between contacts is attained. By proper orientation of the clips, a circular formation of the connecting points is arranged, to agree with the pin positions on the tube base.



Courtesy Electronic Tube Corp.

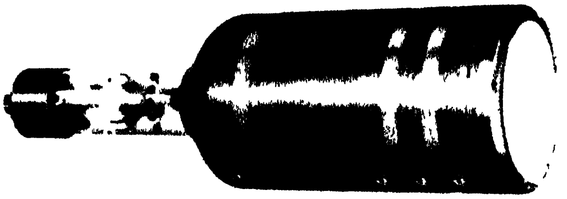
Fig. 4-7.—Twenty-pin tube base and socket.

Location of Electrode Terminals

The location of the electrode terminals is a function of the design of the tube and the preference of the individual manufacturer. It is essential to bear in mind, however, that in the case of special types of tubes, that is, tubes which depart from the conventional single electron-gun construction, there is no standard pattern of electrode terminal location.

In general, low-voltage tube construction is such that all the electrodes terminate on the tube base. In the case of high-voltage tubes, many variations will be found. The accelerating anode usually terminates somewhere along the envelope of the tube beyond the deflection area. The intensifier anode terminal always is found on the bulb, fairly close to the screen. The exception to this arrangement is the tube wherein several such accelerating rings are used; in this case, the terminals are arranged in a row along the length of the envelope. A tube of this type appears in Fig. 4-8.

When thinking about electrodes, we usually have in mind the gun, the second anode, and the intensifier elements. In some tubes, such as the multigun tubes described later in this chapter, intergun shields also have connections; these must be viewed as being a part of the electrode system because frequently they are joined to one or more other tube elements internally, and terminate on the tube envelope.



Courtesy Du Mont Labs

Fig. 4-8.—Recessed-button anode terminals along envelope of 5RP11A tube. Note location of deflection-plate terminals on neck of tube.

Deflection plates terminate in one of two places, depending upon the number of guns and the nature of application. In general, low-voltage tubes used in the conventional cathode-ray oscilloscopes employ such construction that deflection plates terminate at the tube base. This is general practice in the majority of single-gun tubes, even when high voltages are applied to the tube. Although not commercially available, there are some single-gun experimental tubes of domestic make and intended for high-frequency application, wherein the deflection-plate terminals are located on the neck of the tube. Such tubes are commonly in production, however, in foreign countries.

In contrast to the single-gun type of construction, multigun tubes of the electrostatic variety have, as a rule, the deflection-plate terminals placed around the neck of the tube. An example of this is shown in Fig. 4-10B, as well as in several other figures in this chapter. When so mounted, the terminals are located equidistant from the axis of the tube and equidistant from each other. A departure from this form of arrangement is the mounting of the connecting terminals upon an insulating ring which is located around the neck of the tube. This is shown in Fig. 4-15.

A further modification in one multigun tube involves separate terminals around the neck of the tube for the deflection plates and separate terminals around the neck of the tube for the focusing electrodes. A photograph of this tube appears in Fig. 4-4.

Diameter of Electrode Terminals

As has been indicated, tube elements often terminate in contacts which are mounted on the envelope. The important point is that all such terminals are not of like diameter and, in many cases, may require special types of connectors. Reference to the nonuniformity of electrode contacts may appear superfluous in connection with the contents of a book on theory; but the man who has had occasion to work with oscilloscope tubes soon finds out that the inability to make proper connections, because the connecting caps do not fit, can be extremely annoying.

Sometimes these connectors are contained in recesses in the envelope and the snap terminal must fit such a recess; in other cases, the connecting terminal protrudes through the wall, and contact is made by the usual type of grid clip with ceramic insulation on the outside. Also it is significant to note that in those cases where several elements within the tube terminate in connections on the envelope, the high-voltage junctions utilize terminals of greater diameter than the low-voltage junctions.

Multigun Tubes

Many applications demand the simultaneous observation of two or more different phenomena, or at least the presentation of two or more actions, regardless of what may be their nature. These multiple simultaneous presentations may be as simple as two pairs of voltages, or voltages and currents corresponding to actions in four different parts of a system. Or the patterns may be as complex as to require a screen divided into four sections, each of which is excited separately and made to present a scene in a different color, the four individual scenes then being optically formed into a single picture of the combined colors. Such demands are satisfied by multigun tubes.

Whether such multiple traces are developed with electrostatic tubes or with electromagnetic tubes is unimportant at the moment, except that today the majority of multigun tubes available on the open market are in the former category. There has developed during the past two years a definite increase in interest in multiple-trace cathode-ray scopes; at the time of this writing, several manufacturers are producing complete scopes which are capable of displaying as many as five separate phenomena simultaneously. Each of these scopes is described in a later chapter.

The most prominent, although not the only, method of providing multiple traces on the screen of the cathode-ray tube is by means of two or more electron guns contained within the tube envelope. Each gun is complete in itself; it contains its own electron emitter, and if it is of the electrostatic variety, it has a complete set of elements which go to make up whatever specific type of gun it may be. This means that, in addition to the elements which may be located between the control grid and the first or focusing anode, it may have its own second (accelerating) anode, or it may employ an anode which is common to all the guns. Associated with each of these guns is a deflection system consisting of two pairs of deflection plates. As we said earlier, the combination of the gun elements and the deflection plates in a single assembly is a matter of convenience, rather than the acceptance of the idea that the deflection plates are a part of the electron gun.

In the case of the electromagnetic variety of tube which contains two or more guns, each gun conforms with the type normally used in electromagnetic types of tubes. These references to types of guns will become more clear later in this chapter when we examine the fundamental types used in practice, but for the present we can refer to types even though they have not been identified.

As many beams as are active within a multigun tube strike the single screen and produce individual traces which represent whatever may be the nature of the deflection forces the beam experiences while on its way to the screen. By the application of suitable *positioning* potentials, each of these traces can be arranged to appear apart from the others either horizontally or vertically, or both; or any two or all may be so positioned as to appear superimposed upon each other. See Fig. 4-9.

For conventional applications, the domestic commercial market affords multiple-trace scopes which use multigun tubes containing 2, 3, 4, and 5 guns, thus affording a corresponding number of simultaneous traces. Experimental applications have made use of as many as 10 guns within a single envelope. Remarkably enough, this number of guns was used in a tube with a

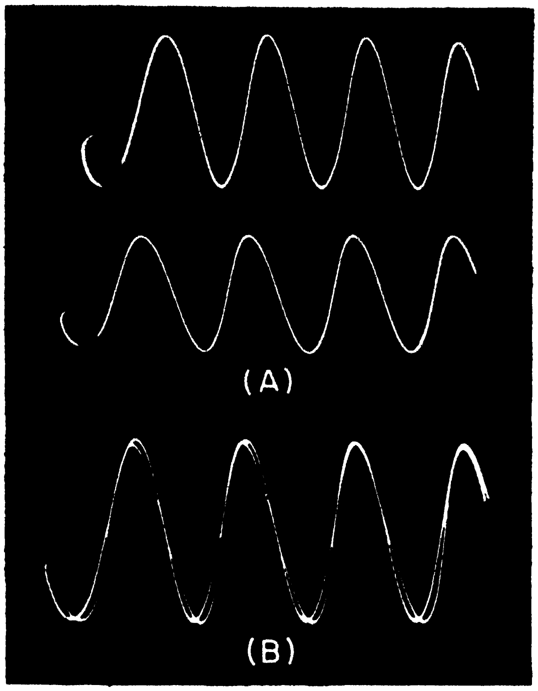


Fig. 4-9.—Oscillograms of traces as they appear on a dual-gun tube. The two traces may be separated, as in (A), or superimposed, as in (B).

7-inch screen; an illustration of the gun is shown in Fig. 4-14. Another interesting design is the use of six guns in a tube which has but a 3-inch screen.

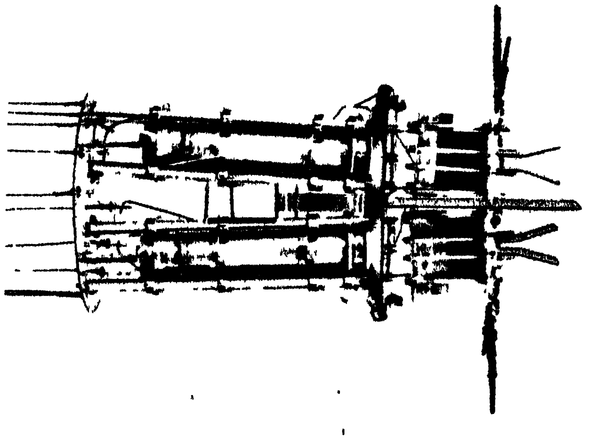
While it is true that the use of two or more guns in a cathode-ray tube, sets these tubes apart from the conventional single-gun variety, the modification in design is limited solely to the gun structure, unless other changes are specifically mentioned. This happens in a few isolated cases, but by and large, it should be understood that the use of two or more guns does not alter the basic manner in which the cathode-ray tube operates. One might say that it corresponds to a car with four forward speeds and two reverse speeds and possibly four-wheel drive, in contrast to the conventional automobile. The basic actions of the different elements in the multigun tube are like the actions of the same parts in the single-gun device, so that the user of a multiple-trace tube can still think in terms of conventional cathode-ray-tube practice.

Arrangement of Multiple Guns

The number of guns used in multigun cathode-ray tubes varies from two to ten. It is possible that more than ten guns have been used within a single envelope, but if so we do not know about it.

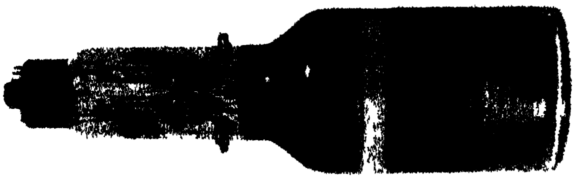
As to the relationship between the screen diameter and the number of guns within a tube, there seem to be almost no limitations, as indicated by the fact that as many as six guns have been accommodated in tubes with a 3-inch screen and as many as ten guns have been used on a tube with a 7-inch screen. Furthermore, as few as two guns may be found in tubes with screens both as small as two inches and as large as twelve inches. As you can see, there is no special relationship between the number of guns within a single envelope and the screen dimension.

Generally, two common gun arrangements are to be found in domestic multigun tubes. The first is that used in dual-gun tubes, wherein the two guns are *in line* along a single horizontal axis. Such an assembly, prior to insertion in the envelope, is shown in Fig. 4-10A and the completed tube is illustrated in Fig. 4-10B. Another example of in-line arrangement is shown in Fig. 4-11, this time a triple-gun assembly.



Courtesy Electronic Tube Corp

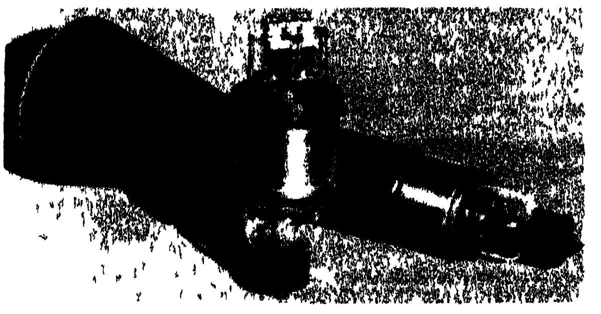
Fig. 4-10A—Dual electron-gun structure of type used in tube of Fig 4-10B.



Courtesy Electronic Tube Corp.

Fig. 4-10B.—5SP11 dual-gun cathode-ray tube.

Several very interesting structural conditions exist in these and other gun assemblies yet to be shown. Electrostatic shielding in the form of plane- or fin-type shields made of nonmagnetic stainless steel is used between the gun assemblies to isolate each from the other. Each shield begins below the control grid and extends upward sufficiently to include the deflection system as-



Courtesy Bell Telephone Labs

Fig 4-11.—Western Electric 330-B triple-gun cathode-ray tube, with gun assembly prior to insertion in envelope.

sociated with the gun. In addition, the shields are mounted in such a manner as to strengthen the mechanical assembly of the guns and to add rigidity.

The effectiveness of the electrostatic shielding is shown by freedom from effect when a strong high-frequency deflection field is applied to one beam and the other beam is not deflected. If the shielding is effective, the nondeflected beam will remain unaffected by the signal applied to the adjacent deflection system.[2]

Another interesting structural condition is the small tilt given each gun relative to the tube axes. This causes the nondeflected beams to strike the screen within a small area. Standard practice is to provide such angular tilt that the two beams strike the screen within a 25-mm square, or a square approximately one inch on each side, with the center of the screen being at the center of the square as shown in solid lines in Fig. 4-12. If the two beams strike anywhere within this area, the structure is within tolerance

Based on personal experience, we find that this tolerance is too great. While it is true that correction voltages can be applied, so as to make these nondeflected beams coincide where they strike the screen, manufacturing tolerances should be decreased so that the two spots from the two nondeflected beams will fall within 5 mm of each other, within an area of 25 sq. mm. instead of 625 sq. mm. The importance of attaining perfect coincidence of the deflected spots is explained in detail later in this text. For the present let it be said that, as a rule, the greater the divergence from coincidence of the two nondeflected beams when they strike the screen, the more difficult is it to properly portray phase relationship between related quantities, when both beams are actuated by a single sweep.

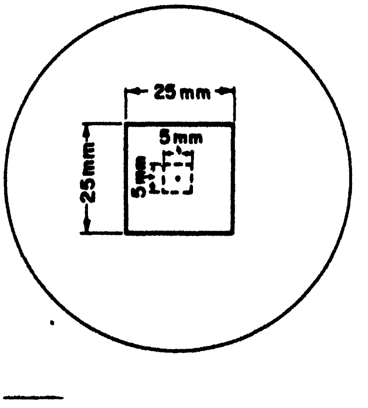
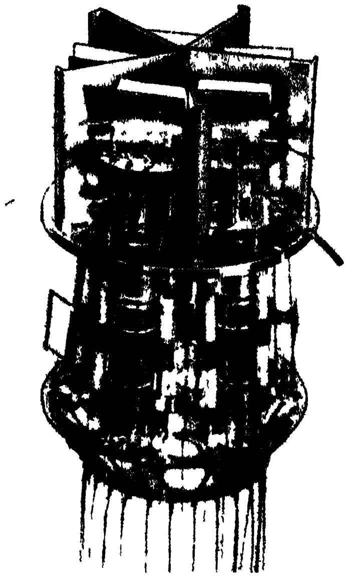


Fig. 4-12.—Manufacturing tolerances permit dual undeflected beams to fall within 25-mm square. Dotted lines show more ideal 5-mm square.

The second arrangement used in multigun systems is circular in pattern. The guns are arranged around the circumference of a circle, being spaced equidistant from the axes of the tube and equidistant from each other around the circumference. An assembly of five electrostatic guns complete with deflection plates, prior to insertion in an envelope, is shown in Fig. 4-13, and a 10-gun assembly is shown in Fig. 4-14. The type of shielding between guns described before is clearly visible in the latter illustration. In the 5-gun assembly, the shielding is limited to the space between the guns.



Courtesy Electronic Tube Corp

Fig. 4-13.—Assembly of five electron guns. Shielding between deflection plates of electron guns is evident.

The wires on the bottom in Figs. 4-13 and 4-14, as well as in Fig. 4-10A, are the connections for all of the heaters, control grids, focusing anodes, and other elements within the guns, whereas the leads on top are joined to the deflection plates. An example of how the deflection plate leads terminate outside the envelope is shown in Fig. 4-15. This is a 6-gun tube. The leads from the gun assemblies can clearly be seen going toward the pins of the tube base, whereas the 24 leads from the six sets of deflection plates terminate in the pins, molded into the insulating ring which surrounds the envelope. The screen diameter is seven inches.

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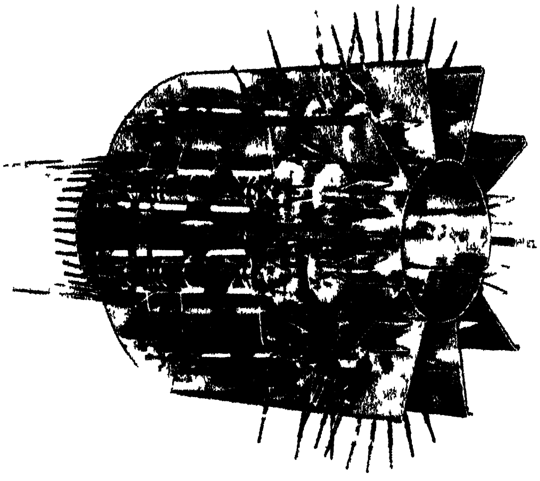
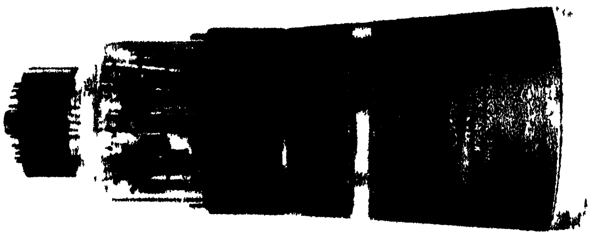
[2] Bamford, H. S., "Multi-gun and Special Purpose Cathode Ray Tubes," paper read before Instrument Society of America, Washington Section, January 19, 1948.



Courtesy Electronic Tube Corp.

Fig 4-14.—Assembly of ten electron guns. Here shielding is used between complete gun structures.

In all of these assemblies, the aim is to keep all the horizontal deflection plates parallel to each other and all the vertical deflection plates parallel to each other. In this way, the divergence of the horizontal traces from parallel paths is kept to a minimum along the length of the trace. The same is true for the vertically oriented traces. The absence of such parallelism constitutes a major problem when it exists — one which tends to impair the usefulness of the tube as a phase indicator for two or more waveforms. Commercially, the parallel alignment of the horizontal traces is kept within 3°. Experience indicates that for schoolroom demonstration of waveform comparison, more than 1° divergence is harmful to the presentation. In fact, ½° variation should be the maximum allowed. This is possible by careful alignment of the elements which make up the gun and horizontal deflection plate assemblies. At this time, when these tubes are being made in comparatively limited quantities, such alignment is costly. But for the good of the art as a whole, the more rapidly it is possible to reduce costs and still provide the parallelism between the time-axis traces, the better it will be for all concerned.



Courtesy Electronic Tube Corp.

Fig. 4-15.—Six-gun 76GAP2 tube, showing unique arrangement of deflection-plate terminals.

So much for the general arrangement of electron guns in multigun tubes. The types we have presented are offered simply as a cross section of current practice in the industry. Without doubt, special shapes of tubes as well as special arrangements of guns are being used experimentally or in some applications which must remain unknown for reasons of national security; even so, the usefulness of what we have shown is not impaired.

The Split-Beam Tube

A method of attaining multiple traces in single-gun electrostatic tubes, aside from electronic switching which will be described later, is by the split-beam method announced by Cossor, a British concern, as early as 1939, and presently employed in the scope offered by this company. This is a dual-trace scope wherein a dual beam is obtained from a single gun by splitting the beam prior to deflection. The process is described during the analysis of guns and deflection systems.

CHAPTER 5

THE ELECTRON GUN

The electric characteristics of the cathode-ray tube involve many different items. Most of them are related to the conditions of use and are specific in terms of operating voltages and current. Still other and equally significant statistics in this category express the operating capabilities of the device.

It stands to reason that these two are closely allied, therefore, any attempt to discuss the subject of electric characteristics must be very broad in its coverage. There is, however, one subject of discussion which seems to be the most important because it has the widest influence upon the operating capabilities and is inseparably tied to the operating voltages. This is the assembly of electrodes known as the electron gun.

Of the major sections of a cathode-ray tube, the electron gun is paramount in determining characteristics, regardless of type of application or purpose of the cathode-ray tube. No matter how the electron beam may be used in the tube, or the significance of the trace on the screen, the function of the electron gun remains unchanged. It is independent of all other actions within the tube, yet is closely related to each, because it is a part of a completely integrated system — the tube as a whole. Therefore, the generalities of its design and operations are vital to the understanding of not only how the cathode-ray tube is used, but also its ability to perform specific functions.

But before discussing the different types of guns, it may be well to devote a few minutes to the review of some fundamentals of vacuum-tube operation. The cathode-ray tube is without question an unusual device, yet in many respects it is comparable in operation to the ordinary vacuum tube. So that we may start from familiar territory to arrive at understanding of the cathode-ray tube, we will review the principles of the triode.

THE TRIODE VACUUM TUBE

In Fig. 5-1 is shown a triode tube using the conventional vacuum-tube symbols. The electron-emitting cathode is designated as *K*, and it is indirectly heated by the heater, which is not shown. The control grid is indicated as *G*, and the plate is marked *P*. Since it is customary in cathode-ray-tube practice to speak about anodes, and inasmuch as it is likewise proper to refer to the plate of an ordinary vacuum tube as an anode, the terms *plate* and *anode* can be used interchangeably in this discussion.

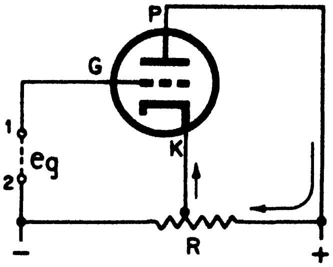


Fig. 5-1.—Conventional triode vacuum tube, showing direction of electron flow.

The power-supply voltage divider is represented by the resistor *R*. The location of the cathode tap along *R* makes the control grid negative and the plate or anode positive with respect to the cathode; this, of course, means that the grid is negative with respect to the plate. The grid circuit contains provision for feeding a signal voltage e_g to the control grid in series with the fixed bias voltage obtained from the power supply. Until this signal is introduced, terminals *1* and *2* are assumed shorted by the dotted-line link. Suppose we begin the discussion by examining several general conditions.

Cathode Emission in the Triode

Let us note that electrons leave the cathode at random velocities; some are emitted at high speeds, ***but by far the majority*** — in fact, we might go so far as to say that ***substantially all*** leave the emitter at ***slow*** speeds. It is possible to say that of their own accord, that is, as the result of the low velocity of emission, the electrons would not travel very far from the cathode. So, without any force of attraction to pull them away from the neighborhood of the emitter, the electrons tend to form a cloud of negative charges around the emitter. Remember that the liberation of electrons by the cathode either in the conventional vacuum tube or in the cathode-ray tube is due to thermal agitation within the emitting material and ***not to the forcible extraction of electrons from the cathode by an external attracting force.***

Many of the emitted electrons return to the cathode, for, as the emitter loses electrons, it becomes progressively more positive, since it contains an ever-increasing number of positive charges. These tend to pull some of the near-by electrons back into the cathode. Moreover, the electrons which surround the cathode are an accumulation of negative charges, and so tend to repel each other and any new electrons which are emitted by the cathode. If nothing but an electron-emitting cathode were contained in an evacuated envelope, there would develop a condition, after emission started, when as many electrons would return to the cathode from the surrounding cloud of charges as left the cathode to join the cloud.

Effect of the Control Grid in the Triode

The control grid is relatively close to the cathode and it is made negative with respect to the emitter. Due to this condition, an electrostatic field exists between the two elements, which, because of the relative polarities, has such direction of action as to prevent electrons from approaching the grid. In other words, the lines of force in this field act in a direction toward the cathode, and the emitted electrons can travel to the grid only if they have sufficient velocity of emission, or if there is an added positive field strong enough to override the negative one. Since the cathode is continually emitting, the space between the grid and cathode becomes full of electrons; we describe this condition by saying that a *space charge* exists between the elements.

Due to the proximity of the control grid to the emitter, a comparatively small difference of potential between the two electrodes can establish a fairly strong field. The intensity of the field is a function of the voltage effective on the grid. With e_g equal to zero, the voltage effective on the grid is the bias voltage obtained from the power supply. But if e_g is not zero, the effective voltage is either greater or smaller than the bias voltage by an amount determined by the magnitude of e_g and its polarity relative to the bias voltage.

The more negative the control grid relative to the cathode, the stronger is the repelling field and the fewer the number of electrons which approach the grid. This number may be made so small by means of suitable values of grid voltage, that for all practical purposes, the movement of electrons to and through the grid may be considered zero. Due to the random velocity of the emitted charges, some will get to the grid, for it is in the path of the electrons advancing toward the highly positive anode or plate, which will be discussed shortly. This is so even when the grid is negative by a certain amount. The less negative the grid, the lower the intensity of the repelling field, and whether or not electrons can advance past the grid depends on the effect of the plate as an attracting influence.

The reference to electrons reaching the grid is made for a definite purpose. If, during the operation of the tube, the grid is made somewhat negative relative to the cathode, the voltage applied to the plate will cause the movement of charges past the grid. Some electrons will strike the grid (just as some raindrops may strike a wire fence, although most pass through) resulting in a small grid current. This happens, remember, even when the grid is not positive — its significance is that some movement of charges will take place, and if the circuit resistance is high, which is not the case in the example being discussed, the resulting voltage drop across the grid-circuit resistance will affect the bias. Therefore, the matter of grid-circuit resistance is important; not in the circuit we are discussing, but because of circumstances to which we will refer later, and because it explains in part why cathode-ray-tube specifications refer to maximum values of grid-circuit resistance. This will be mentioned again later, when the fundamental electrostatic type of gun is discussed.

Effect of the Plate in the Triode

The tube envelope also contains the plate, also referred to as the anode. This electrode is made very positive with respect to the cathode and control grid; an electrostatic field, therefore, exists between the plate and grid and between the plate and cathode. The plate, being positive relative to the cathode and grid, can exert an attraction upon the charges in the space cloud, tending to pull these charges, through the meshes of the grid, to itself. In other words, the attracting influence of the voltage on the plate can accelerate electrons to that element; such movement of charges is called plate current.

The effectiveness of the plate in causing the movement of charges depends upon numerous variables, some of which are electrical in character (the operating voltages) and some of which are determined by the geometry of the system. These are related to each other; but since the tube geometry is fixed, we need discuss only the electrical variables. Only these are within control of the operator.

If, for the moment, we neglect the presence of the control grid and view the envelope as containing only the emitter and the anode, the magnitude of the plate current, that is, the ***number of charges*** which advance to the plate, is strictly a function of the difference of potential between the plate and cathode. The greater the voltage applied to the plate, the greater the number of charges attracted to the positively charged electrode, and the higher is the plate current. (There is a limit, when *all* the emitted electrons are attracted to the anode, leaving no space-charge cloud; this limit occurs well beyond the maximum operating value of plate current.) Still another effect, and one which is pertinent to the cathode-ray-tube operation, even if not especially significant in ordinary vacuum-tube practice, is the velocity of the electrons which constitute the plate current. *The higher the plate voltage, the higher is the terminal velocity of the electrons when they reach the plate* — which means the higher velocity of the electrons during their travel between the space charge and the plate. The basis for this was discussed in Chapter 2.

Unfortunately, the charges do not advance to the plate under the influence of the tube plate voltage as a ray of electrons, having no cross-sectional size. The shape of the tube electrodes gives rise to a stream of charges of substantial cross section, as determined by the dimensions of the emitting surface as well as the dimensions of the plate which the electrons strike. Just what we mean is shown in simplified form in Fig. 5-2. Despite this type of distribution of the charges in the plate current, it is still possible to view the stream of electrons as having charge *density* per unit cross-section area, so that an increase in plate current corresponds to an increase in the electron density.

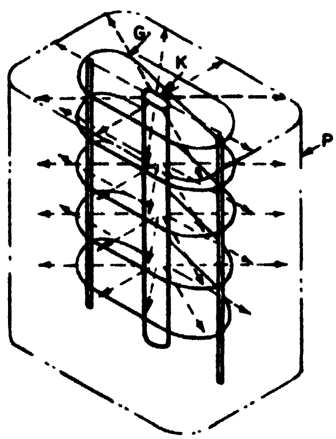


Fig. 5-2.—Electron emission from cathode of triode vacuum tube.

Effect of the Grid and Plate

Let us now recognize the presence of the grid and analyze the effect when both plate and grid are active. We said that when these two electrodes are subjected to operating voltages, two fields exist in the tube. We also said in a previous chapter that lines of force cannot cross each other, so that whatever may be the individual operating voltages applied to these two electrodes, there comes into being within the tube a composite field which embraces the effects of the grid and the plate.

With the grid negative with respect to the cathode, and closer to it than is the positively charged plate, the effect of the grid upon the electrons in the space charge is greater than that of the plate. For equal changes in potential between the cathode and each of the other elements, the effect of the grid is greater by a multiple corresponding to the amplification factor of the tube; this factor is determined by the tube's structural geometry. In other words, a change of one volt on the grid may have as much effect upon the movement of charges to the plate as a change of 20 volts in plate voltage. Thus by the application of a suitable negative voltage to the control grid, it is possible to completely nullify the attracting influence of the plate, regardless of the value of plate voltage. Under such conditions, there is no movement of electrons beyond the grid, consequently, there is no plate current. The tube is then said to be at *cutoff*.

The amount of negative grid voltage required to stop the flow of plate current is known as cutoff grid voltage. For every value of plate voltage within the usable range, as established by the tube manufacturer, there is a value of grid bias which will cause plate-current cutoff. Obviously, for any value of plate voltage a reduction of the grid bias below the cutoff value, that is, making the grid less negative, will cause an increase in the plate current. Making the control grid more negative than the cutoff value will have no effect since plate-current flow has already stopped.

The electric relationship between control-grid voltage and plate current for one or more values of plate voltage can be shown graphically, as in Fig. 5-3. The curves illustrate not only the change in plate current due to changes in grid voltage but also the required grid voltage above cutoff to cause a specified plate current. Such a family of curves is very important to the designer of equipment. To us it is important because it is the type of curve we will discuss later in connection with the characteristics of the control grid in the electron gun.

Several very interesting comments may be made concerning the E_g-I_p family of curves. First of all, such a family of curves is not peculiar to triode vacuum tubes; as we said before, it resembles very closely the *grid-drive* characteristics of the control grid in the

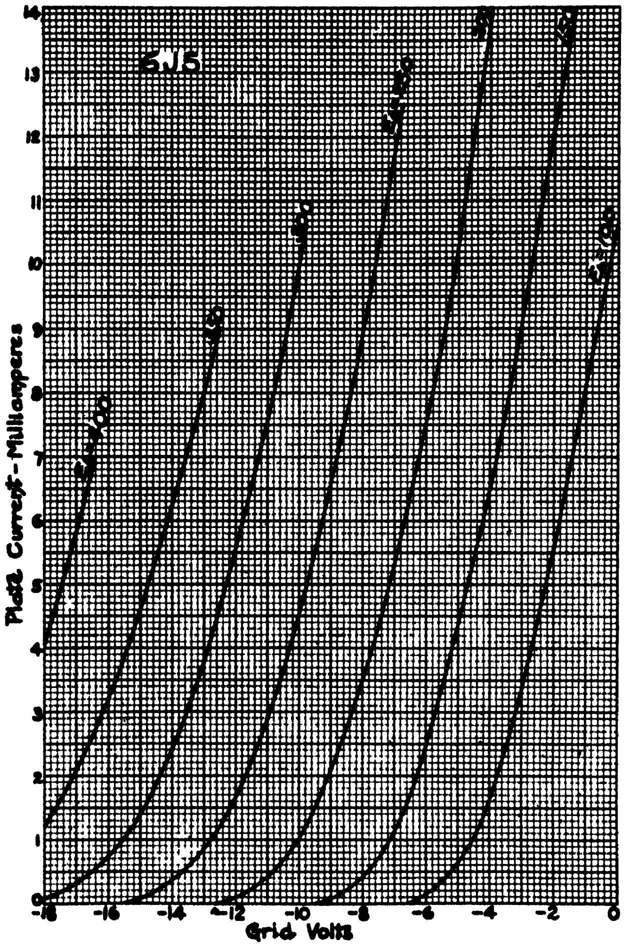


Fig. 5-3.—E_g-I_p family of curves for 6J5 triode.

electron gun. As will be seen later, the functions of the control grid in the electron gun parallel closely the functions of the control grid in the ordinary tube.

Selecting one of the curves, for example that which corresponds to $E_b = 100$ volts, a grid bias of -6.5 volts causes plate-current cutoff, whereas a bias of -4 volts (2 volts above cutoff) raises the plate current to approximately 1.5 milliamperes, and with gradually decreasing values of negative grid voltage, the plate current rises to about 10.6 milliamperes at zero bias. Like manner of behavior, except for differences in the actual values of grid voltage and plate current, is experienced with higher values of E_b; it is evident that the higher the plate voltage, the higher must be the negative grid voltage to produce plate-current cutoff in any one particular type of tube.

Let us examine this action from a different viewpoint. What do we mean when we talk about various values of plate current, or plate-current increase or decrease, or plate-current cutoff? Fundamentally, we are speaking about the ***number of electrons*** which are permitted to pass through the meshes of the grid and advance to the plate. Although we use units of current, rather than actual electron count to express the plate current, it is nevertheless proper to think of the plate current as comprising great numbers of individual electrons. Thus no matter what the shape of the stream of charges which advances to the plate, whether a concentrated beam or a relatively broad and deep stream, changes in the number of electrons are changes in density. *A cardinal function of the control grid, therefore, is the control of the number of electrons* in the plate or anode current, or in the density of the charge stream.

There may seem to be a conflict between the function of the control grid and the plate because both electrodes are associated with the action of controlling the number of electrons which form the plate current. Actually no such conflict exists. Both electrodes have an effect on the plate current, therefore, both display an influence on the number of electrons which advance to the plate. However, in this respect, the control grid is the main contributor, as demonstrated by the fact that the magnitude of its control is very much greater than that of the plate. In certain types of tube, such as pentodes, the effect of the plate voltage on the value of plate current is extremely small, whereas the action of the control grid remains very great. Although not quite accurately (because each electrode affects the other to some extent), we can generally say that the *control grid* determines the ***number of charges*** which form the plate current, and the ***plate*** determines the *velocity* of these charges as they move to the plate. It so happens that during the process of ordinary vacuum-tube operation, electron velocity receives very little attention, except from the tube designer. The user has little interest in it.

In the cathode-ray tube, the conditions are different; electron velocity is important in understanding the operation of the tube. Changes in plate current have a different meaning. Now it is a matter of beam velocity dimensions and trace brilliance, or stated differently, light output.

Fig. 5-3, it should be noted, indicates no positive values of grid voltage. While *this grid is always negative with respect to the cathode in the case of the present cathode-ray-tube designs,* there are a few vacuum tubes which are designed to run at zero or slightly positive grid bias.

Signals on the Control Grid

Up to this point, the grid was maintained at a certain potential relative to the cathode by means of the grid bias developed across a portion of *R* in Fig. 5-1. Let us now examine what happens if a voltage e_g is injected in series with the grid bias at terminals *1* and *2* in Fig. 5-1. Understandably, the signal voltage e_g may have either positive or negative polarity and it may be of variable amplitude. Being in series with the grid bias, it may add to the bias or it may buck the bias. In either case, it changes the effective voltage at the grid, and in that way determines the effectiveness of the plate voltage as well as the space charge.

If we apply a sine wave of voltage to the grid, the grid voltage will vary sinusoidally around a mean value established by the fixed bias, and if the value of the signal voltage is within prescribed limits, the plate-current changes will faithfully follow the changes in grid voltage. This is the equivalent of linear amplification as illustrated in Fig. 5-4.

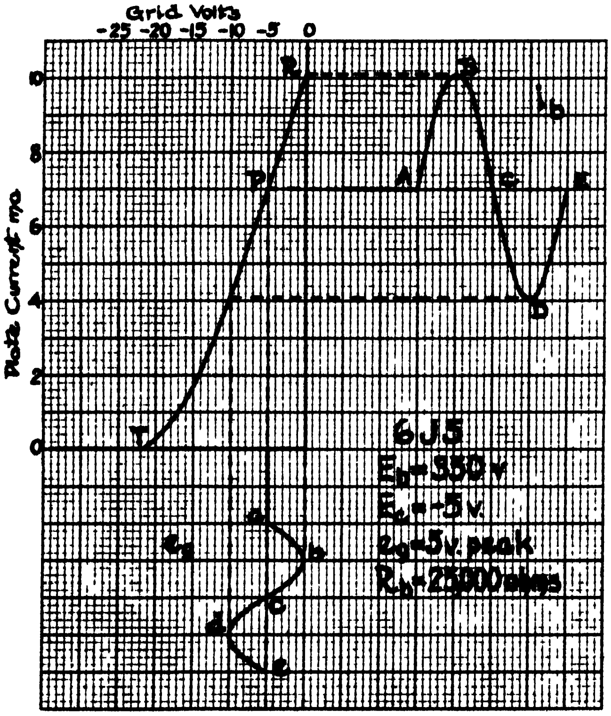


Fig. 5-4.—Signal voltage applied to transfer characteristic of 6J5 results in amplified variation of plate current.

There is correspondence between the vacuum tube and the cathode-ray tube, but the action of signal amplification is not one of them. When analyzing the action of the ordinary vacuum tube, we normally concern ourselves with the waveshape of the plate current relative to the waveshape of the grid voltage. In Fig. 5-4, the plate current is a faithful reproduction of the voltage effective at the grid. In cathode-ray-tube work, however, we are concerned with the *change in density of the beam current* as the result of changes in grid voltage. Such changes may reduce the beam current sufficiently so that the luminosity of the spot on the screen is reduced to zero, or there may be such increase from zero beam current as to produce a bright spot.

Perhaps this groundwork can be better appreciated by reference to Fig. 5-5. Here normal operation is beyond plate-current cutoff, thus making the tube inoperative until the grid voltage becomes less negative (but never positive) and allows a flow of plate current. The operating point in the grid circuit is established by the fixed bias, which as can be seen, is beyond the cutoff value by an amount which offsets a portion of the positive signal voltage. During most of the positive half of the signal cycle, the tube remains cut off; then as the positive peak of the input signal is approached, the cutoff limit is exceeded and a short period of plate-current flow ensues. To say the least, the shape of the curve of plate-current changes is very much unlike the shape of the voltage wave on the grid, but the desired condition of momentary plate-current flow has been achieved.

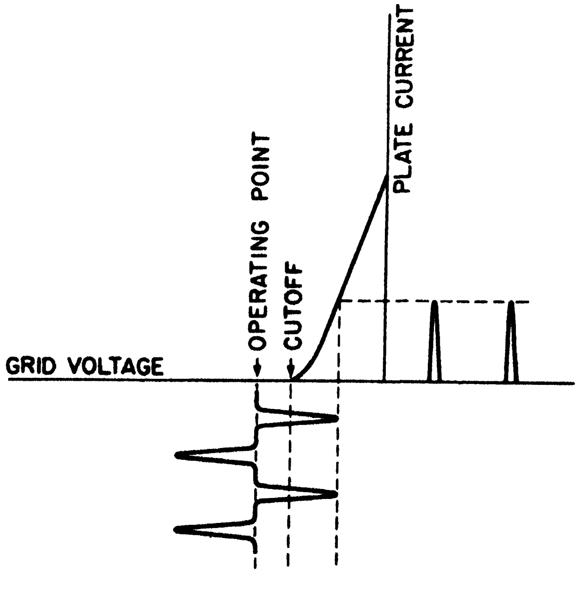


Fig. 5-5.—Peaked waveform applied to triode biased beyond cutoff results in plate-current flow of short pulses.

The negative half-cycles of the input signal voltage have no effect upon the plate current; all they can do is make the grid more negative than it was with zero signal voltage. Since plate current was already cut off with zero signal voltage, a more negative grid contributes nothing to the over-all action.

The possible uses of such a circuit arrangement are numerous, as are the possible shapes of the input signal voltage. The amount of change in plate current, during the moments when the grid voltage is less than the cutoff value, may be any value; determined by the circuit, the operating voltages, and the tube in question. All of these conditions are incidental to the primary action wherein, by virtue of a certain shape and amplitude of the grid voltage, the flow of plate current can be cut off for a specified period of time and be permitted to exist for another specified period. Moreover, the *value* of the current during its existence — or the electron density of the current — likewise may be established as desired.

By modifying the bias and the waveshape of the control voltage applied to the grid, the characteristics of the plate current are changed, as in Fig. 5-6. Instead of having short-duration peaks, the control voltage is a square wave. The rise in plate current is sudden, even more so than in Fig. 5-5, but now the *period of current flow* is longer. Here too, only the positive half-cycles are effective, but unlike the previous mode of operation, the grid bias is selected so that zero signal voltage coincides with plate-current cutoff. Plate current flows during the *entire* positive half-cycle, and has a period substantially that of the time *t* of the positive half-cycle of grid voltage. As before, the negative half-cycles contribute nothing to the plate current, other than to keep the tube at cutoff.

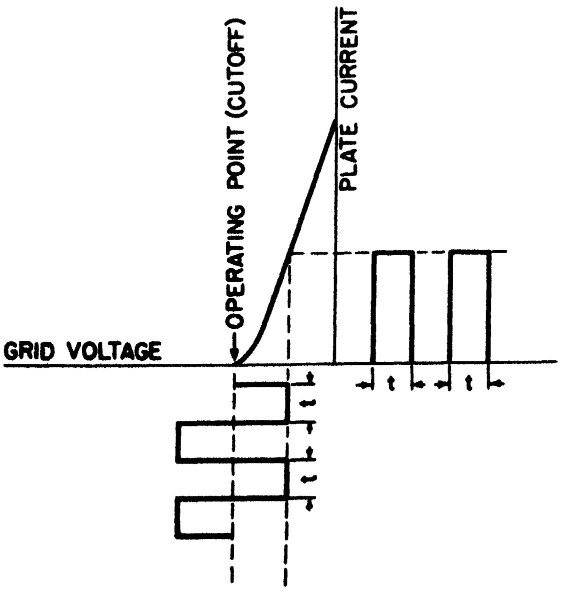


Fig. 5-6.—Square wave applied to triode biased at cutoff results in square-wave plate-current flow.

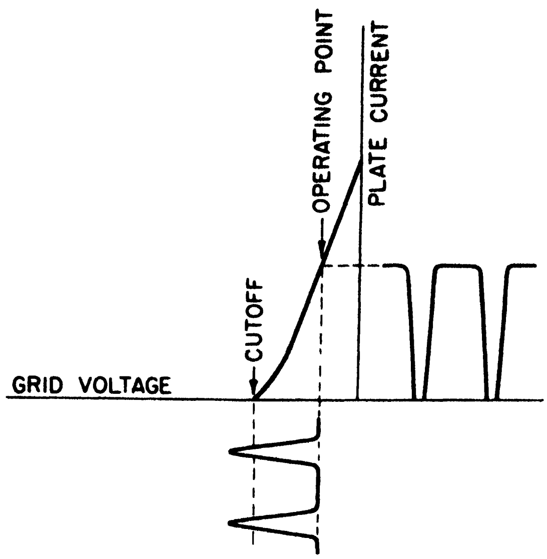


Fig. 5-7.—Negative pulses applied to triode biased at linear portion of transfer characteristic, and having sufficient amplitude to drive grid beyond cutoff, result in distorted plate-current reproduction.

Whether the signal voltage applied to the control grid is sharply peaked and intended to cause an increase in plate current, or the circuit is so arranged that only the negative half-cycle will be effective by driving the tube to cutoff or beyond, as in Fig. 5-7, is immaterial. That is purely a matter of circuit design and the goal desired. The point is that the control grid can be made to control the value of plate current and the duration of plate-current flow.

Modulation of the Plate Current

The effect of control-grid voltage variations upon the plate current is occasionally referred to as *modulation.* This is an extension of the term, as ordinarily used in radio, wherein one characteristic (such as frequency or amplitude) of a carrier wave is varied in accordance with an applied signal. In the present case, it is the normally constant current which varies according to the applied grid signal.

Modulation of the plate current can also be accomplished by changing the plate voltage, in which case, the variations in plate current will conform with the character of the variations in plate voltage. Thus, if a source of voltage, having frequency *f* were placed in series with a fixed d-c plate-voltage supply, the tube plate current would vary in amplitude at a rate corresponding to the frequency *f*. The two kinds of modulation described above are called grid modulation and plate modulation, respectively.

The discussion of the triode just given and the comparison with cathode-ray tubes by no means covers all similarities between the cathode-ray tube and the usual vacuum tube. There are analogies for several other parts and actions, such as the screen grid, but the analysis of the cathode-ray tube may be readily understood without further comparison.

ELECTRON GUNS IN ELECTROSTATIC TUBES

In order to present, in the most logical order, the details about electron guns, the subject will be divided into two general parts. The first will deal with electrostatic types and the second part will analyze the varieties employed in electromagnetic-type cathode-ray tubes. The only deviation from the organization indicated by these remarks is the inclusion of the gas-focused type of cathode-ray tube — the first practical forerunner of cathode-ray tubes used in radio and allied research — among the electrostatic electron guns.

The Basic Electron Gun

For our purposes, the term "basic," as applied in this paragraph heading, signifies that the gun is being considered in its crudest form, being sufficient only to show the most important principles. Practically, the basic gun would be quite unsuited for use in a tube.

Such a basic gun would consist of an electron-emitting cathode *K*, heated either directly or indirectly. The tube also would contain an anode *A* in the form of a simple disk with an aperture at its center, and a screen which will fluoresce when struck by electrons. The entire assembly is contained in an evacuated envelope with suitable connecting means for applying the required potentials to the different elements. These potentials would be available from a power-supply source whose voltage divider we will identify as *R*. An idea of this arrangement may be had by reference to Fig. 5-8.

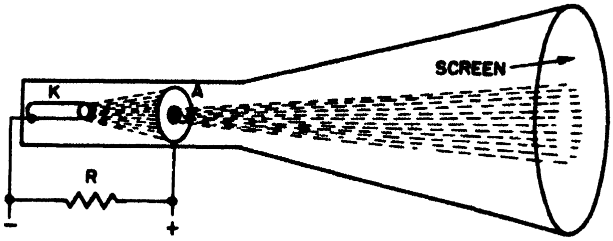


Fig. 5-8.—Elementary electron gun consisting of cathode and anode.

The anode, being made positive relative to the cathode, will accelerate the emitted electrons toward itself, with a velocity determined by the difference of potential between the two elements. Because of the construction of the system, the majority of the electrons drawn toward the anode will strike that surface, but a few will pass through the aperture, and due to the velocity acquired under the influence of the anode voltage, will continue on their way and eventually strike the screen. Depending on the electron velocity and the number of electrons, they may or may not cause fluorescence. Those electrons which leave the cathode parallel to and close to the axis will find their way through the opening and perhaps strike the screen at its center. Those which leave the cathode at small angles to the axis also may find their way through the anode opening; but because of the angle they made with the axis when they left the cathode, their paths will diverge from the axis as they travel the distance between the cathode and the screen, and they will strike the screen at various distances from the center.

Such an arrangement obviously is ineffective for a number of reasons. Even under ideal conditions, the smallest diameter of the beam would be equal to the diameter of the anode opening. If this were made very small so that the undeflected spot on the screen would be small, the density of such a beam would be totally inadequate because there is no concentration or focusing. In the meantime, the emission from the cathode would be wasted almost completely, because without any space-charge control between the cathode and plate, the entire cathode surface is emitting to form the anode current.

If the anode opening were made large in comparison with the diameter of the anode disk, then the diameter of the beam would be far too great to be of any use, especially if the mutual repulsion between the charges could spread them apart. Most certainly such a wide beam would be of little use for deflection purposes because the trace it would produce on the screen would be wide and indistinct. In general, such a system is incompatible with modern needs.

Although quite impractical, this system has the merit of illustrating a few of the principles which are the basis of the modern electron guns. For example, the plate current in the conventional vacuum tube, which contains an electron emitter and a plate, corresponds to the current which passes between the cathode and the anode as indicated by the arrow. But some of the charges do not terminate on the anode, but travel to the screen, so that a new designation for this current is required. This is the *screen current,* or *beam current,* although our personal feeling is that the beam current should be considered the aggregate of all the charges which leave the cathode for travel through the system. For the present, our terminology will be sure that the charges which advance to the screen constitute the screen current.

What happens to these charges? At one time it was thought that they leaked off the screen and returned to a point of positive potential, in this case the anode. That, however, has been questioned, and the theory now accepted is that the screen acts as an emitter of secondary electrons as the result of the bombardment of the surface by the electron beam. The amount of such secondary emission is proportional to the number and velocity of the electrons which strike it. These secondary electrons then are attracted to the nearest point of positive potential. In this connection, let it be understood that any subsequent references to the movement of charges from the screen to electrodes in the tube refer to the secondarily emitted charges.

Whether the screen current is high in comparison with the anode current is simply a matter of how many electrons, comparatively speaking, fall upon the anode as compared with those striking the screen. If the latter are few in number, whereas the former are very many, then the screen current is small and the tube current is almost entirely the anode current.

If the number of electrons striking the screen far exceeded the number striking the anode because the opening in the anode was very large, then the current to the anode would be substantially the screen current. This may be difficult to envision in the example being considered, but it is nevertheless an important point, because of the structure of certain cathode-ray tubes. In these types, an element, acting as a "post-deflection" accelerating anode, does not attract electrons out of the beam, but it does receive the electrons which strike the screen, or rather are secondarily emitted. Thus, the measurement of the current in the post-deflection accelerating circuit is in reality a measurement of the screen current; for electrical equilibrium, the number of electrons leaving the screen must be exactly equal to the number arriving.

Practically, the velocity of the electrons is determined solely by the voltage applied to the anode. The temperature of the cathode may determine how many electrons are available for acceleration by the positive voltage on the plate, but since the usual emission from a heated cathode is at very low velocity, it is the attracting influence of the voltage on the plate which makes them move across a space. The higher this voltage the faster they will move; likewise, the greater will be the number of electrons attracted to the anode per unit time, so that the current will be greater. It is the velocity, however, which interests us most. Should the electrons lack adequate velocity because of low anode voltage, they still may be accelerated by the positively charged anode, and may travel to the screen, but they will not excite the coating and cause fluorescence. On the other hand, lacking an adequate number of electrons, even at high velocity, the beam will not excite the screen sufficiently to be useful. So we see that cathode temperature and anode voltage influence the anode and screen currents; that anode voltage influences the beam velocity and, in that way, the excitation of the screen.

The Electron Gun in the Gas-Focused Tube

Present-day design of electron guns is a great departure from the very simple gun in the gas-focused cathode-ray tube introduced by J. B. Johnson in 1922.[1]

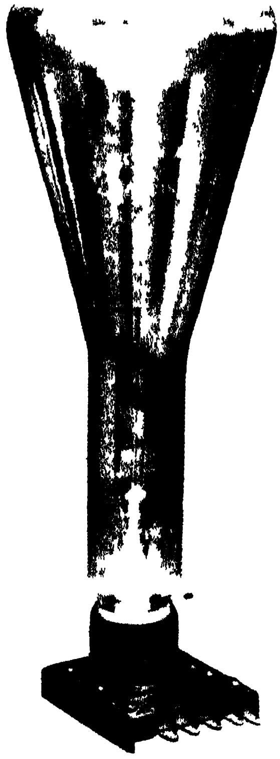


Fig 5-9.—Western Electric 224 gas-focused cathode-ray tube.

Courtesy Bell Telephone Labs.

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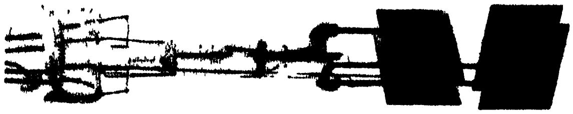
[1] Johnson, J. B., "A Low Voltage Cathode Ray Oscillograph," *Bell Sys. Tech. Jour.*, p. 142, vol. 1, November 1922.

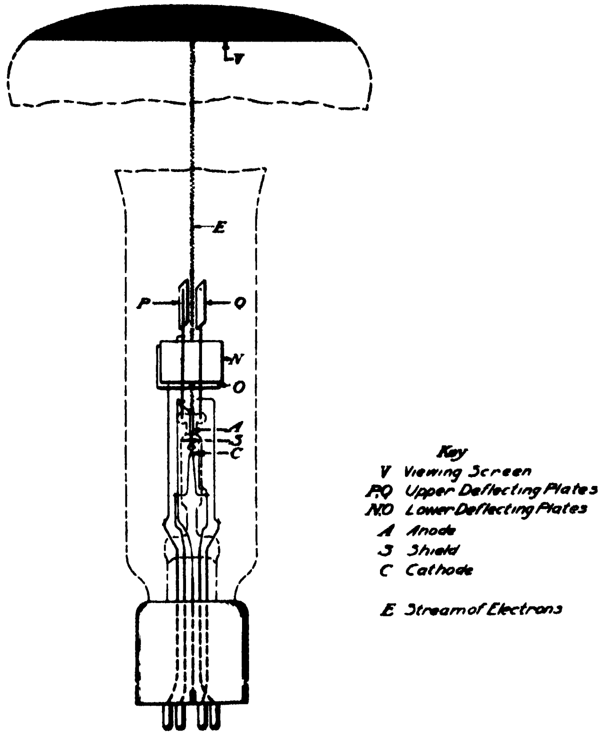
Yet the Western Electric 224 cathode-ray tube was the sensation of its time in research circles. It was the first cathode-ray tube used by the writer, and after two years of experimentation, we described a cathode-ray oscillograph for use in the maintenance branch of the radio industry.[2]

In appearance the tube resembled modern types, as is shown in Fig. 5-9. Although preceding today's envelopes by almost three decades, its influence in design cannot be denied. The assembly of components comprising the tube, with the exception of the screen, is shown in Fig. 5-10; the screen, as in present-day equipment, was painted on the inside surface of the wide portion of the tube. The equivalent schematic representation is shown in Fig. 5-11, which we shall use for the explanation of the tube and its gun.



Courtesy Bell Telephone Labs.

Fig. 5-10.—Electron gun of Western Electric 224 gas-focused cathode-ray tube.



Courtesy Bell Telephone Labs

Fig. 5-11.—Schematic representation of Western Electric 224 gas-focused cathode-ray tube.

Among the numerous points of difference between this tube and those in use today, was the presence of an inert gas in the main envelope as a means of focusing the beam. This alone sets the tube apart from those in use today. In fact, ever since about 1930, virtually all cathode-ray tubes which have found application in research, television, and other fields have been of the high-vacuum variety with either electrostatic or electromagnetic focusing.

Since we are concerned mainly with the electron gun, let us describe it, making reference to Fig. 5-11. The cathode *C* was an oxide-coated platinum ribbon, similar to the filaments used in the famous "E" and "J" type vacuum tubes, which made the name of Western Electric so well known to radio amateurs shortly after World War I. The anode *A* was a thin platinum tube one centimeter long and one millimeter in diameter, one end of which was about one millimeter from the top of the cathode filament loop. The other end was directed into the main envelope toward the deflection plates and fluorescent screen. Between the anode and the cathode there was a metal shield *S*, with a small opening through which the emitted electrons passed to the anode opening.

The anode, connected to a positive voltage source, accelerated the emitted electrons toward itself. Most of them entered the anode tube, but only a comparatively few passed through the full length of the electrode, eventually to form the electron beam in the main envelope. The electrons which left the anode in the main envelope were not concentrated into a dense beam, due to the comparatively low velocity and because of mutual repulsion between the individual charges. How these charges were focused into a beam will be explained shortly.

The complete assembly of the cathode, shield, and tubular anode was sealed within a glass housing, located above the glass stem, as shown in Fig. 5-10. The anode protruded through the sealed glass housing into the main envelope. These parts constituted the complete gun assembly.

The use of gas for focusing is an interesting phenomenon, which depends upon the difference in mobility of electrons and positive ions. A positive ion is an atom which is deficient in electrons, and therefore, bears a preponderance of positive charges. After evacuation of the envelope so that a high vacuum existed, a small amount of gas such as argon, at a pressure of a few thousandths of a millimeter of mercury, was introduced. At this gas pressure, the dimension of the spot which appeared upon the screen was controllable to about one millimeter in diameter. Without this gas, that is, with just the high vacuum in the tube, the dimension of the beam incident on the screen would have produced a luminous spot about one centimeter in diameter, or ten times as wide.

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[2] Rider, J. F., "The Cathode Ray Oscillograph Tube," *Radio Engineering*, pp. 982-986, vol. 7, October 1927.

Because of the action of the positive voltage applied to the anode, an initial, though not concentrated, beam of electrons was available from the opening of the anode. In the absence of gas in the tube, the mutual repulsion between these slow-moving electrons (about 6,000 miles per second) would be sufficient, during the time required to travel from the anode to the screen, to spread the beam sufficiently to make the tube useless. The argon gas present in the envelope counteracted the tendency toward spreading of the electron stream by being active in forming an electrostatic field, radially directed toward the line of travel of the beam electrons.

This radial field was the result of excess of positive charges in the electron stream and an excess of negative charges in the space outside the beam. Such a distribution of charges was produced by some of the electrons of the stream passing through the gas and colliding with gas molecules, thus ionizing them by dislodging electrons from the atoms. Both the colliding electrons and the secondary electrons left the beam, but the heavy positive ions received very little velocity from the impact and drifted out of the beam with only their own comparatively low thermal velocity. Positive ions, therefore, accumulated along the length of the stream and exceeded in number the passing negative charges. At the same time, electrons were moving at random outside the stream, producing a negative charge in that area. There then developed a field surrounding the stream which tended to pull the electrons inward. If there were only the mutual repulsion between the electrons for which to compensate, this would be done when the number of positive ions in the beam equaled the number of electrons. But there was, in addition, the original divergence of the beam which had to be overcome. This was successfully accomplished by arranging that for the gas pressure used, a certain electron emission occurred, which was controllable by means of the filament rheostat.

To prevent damage to the cathode by positive-ion bombardment, the emitting portion of the filament was formed into a loop which had a diameter somewhat greater than the diameter of the aperture in the shield. In this way, the sensitive emitting surface was kept out of the path of the positive ions, which, as they drifted down the anode opening, would move past the active part of the filament. Thus, if they did strike the filament, they did so at an inactive point.

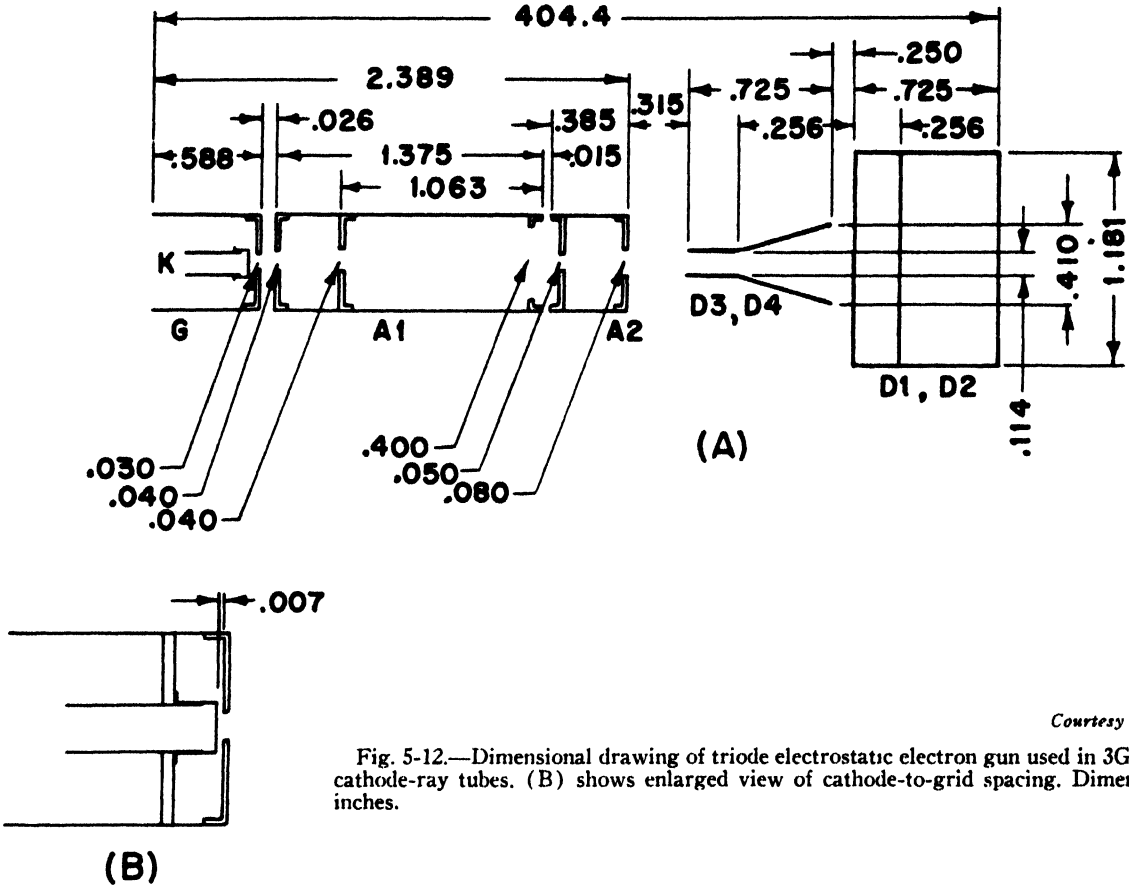
Obviously, the gas was doing that which presently is being accomplished by means of electron lenses, and the radial field previously mentioned was, in fact, an electrostatic field. The only variable control which was found in the gas-focused tube system was the rheostat that controlled the cathode temperature. The adjustment of the cathode emission served a dual purpose, varying the intensity of the spot as well as its focus. Both were accomplished simultaneously because there was a critical emission level at which the stream was most dense, yet focused at the distance where the screen was located. Increasing the emission beyond the critical point for a certain tube geometry resulted in bringing the electrons to a focus in a shorter distance.

Of interest, also, was the frequency limit imposed on the utility of the tube because of the time required for proper ionization. Experiment disclosed that this time was about one microsecond, so that the frequency of the deflection voltages was limited to less than 1,000,000 cps. At this frequency, fuzziness of the trace developed, whereas at about 100,000 cps or slightly higher, the trace was still sharp.

The Triode Electron Gun

The *triode* electron gun is not the simplest used in practice, but it was the standard type in electrostatic cathode-ray tubes for many years, up to and including the recent war years. It has been replaced by modified versions, but it still can serve as a means of comparison. In fact, some of the tubes which employed it are still available so that it justifies description, if only on that account. It is the type of gun usually found in early cathode-ray tubes which bear type numbers without the suffix letter *A*. As an example, the 902 employs a triode gun, whereas the 902-A employs a different type which will be described subsequently.

The triode gun is an assembly of electrodes as shown in Fig. 5-12A. This is a typical assembly showing relative positions of the elements along with their dimensions and those of the apertures and element spacing. The gun consists of an indirectly heated cathode *K*, which is in the form of a cylinder closed off at one end by means of a plate. This plate is coated with barium or strontium oxide, materials which are profuse emitters of electrons at comparatively low tem-



Courtesy Du Mont Labs

Fig. 5-12.—Dimensional drawing of triode electrostatic electron gun used in 3GP- and 5LP-cathode-ray tubes. (B) shows enlarged view of cathode-to-grid spacing. Dimensions are in inches.

peratures. It is brought to the emitting temperature by means of a heater element; this is not shown, but is contained within the cathode cylinder and separated from it by a heat-conducting ceramic sleeve.

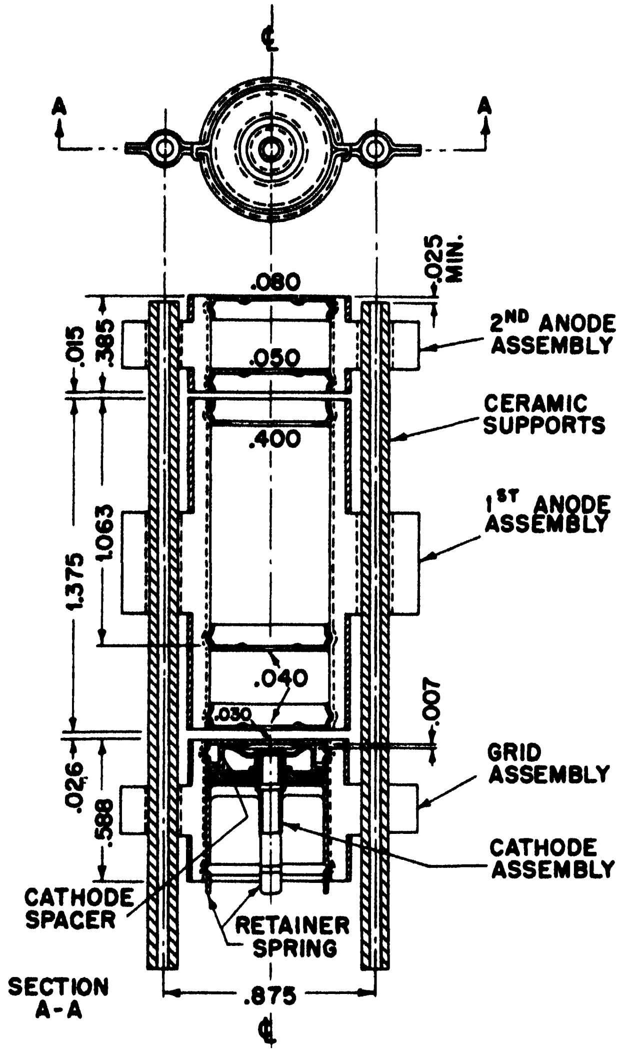
Surrounding the cathode is a cylinder *G* which has a baffle containing a tiny aperture at its center. While this cylinder as a whole is normally identified as the control grid, specifically, the baffle with its opening is the actual control grid. The remainder of the grid cylinder acts as a shield around the cathode. The grid aperture is smaller than the emitting surface, and the spacing between the control-grid aperture and the cathode emitting surface likewise is very small. A detail drawing of the cathode-to-grid spacing is shown in Fig. 5-12B.

Facing the control grid is the first-anode cylinder, designated *A1*. Physically, it is coaxially symmetrical with the control-grid cylinder and contains several baffles, each with an aperture at its center. This electrode, in conjunction with the control grid, forms the cathode lens, sometimes referred to as the first focusing lens. It is responsible for the formation of the first cross-over point.

Adjacent to anode 1 is another cylinder designated *A2*, the second or accelerating anode. This structure is coaxially symmetrical with anode 1 and has two baffles, each with an aperture at its center. An extension of anode 2 exists in the form of a conducting coating on the inside of the envelope, extending from the approximate limit of the anode cylinder in the gun almost to the screen of the tube. This arrangement is standard practice in electrostatic tubes. The second-anode cylinder and the extension on the inside of the envelope are connected to each other by means of a "spider" arrangement of electrically conducting spring contacts which are mounted on the gun assembly.

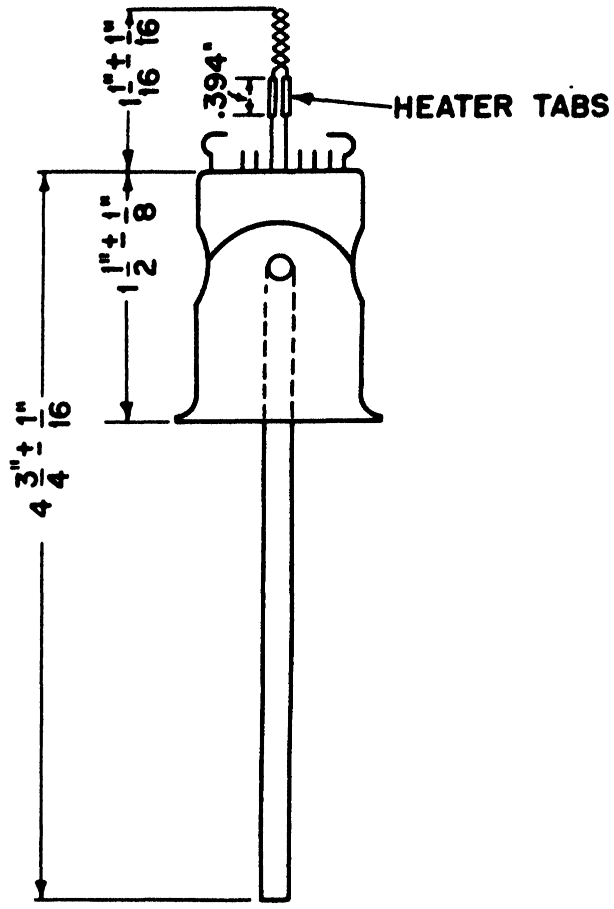
Anode 2 serves a dual purpose. It is the accelerating anode whereby the final velocity is imparted to the emitted electrons so that they will traverse the entire system and advance to the screen; and the combination of anodes 1 and 2, at that point where they face each other, constitutes the principal focusing lens.

The deflection plates, *D1-D2* and *D3-D4*, are also shown on the drawing of Fig. 5-12A. Strictly speaking, these elements are not part of the electron gun. However, because of manufacturing convenience, they



Courtesy Du Mont Labs.

Fig. 5-13.—Sectional view of triode electrostatic electron gun. Dimensions are in inches.



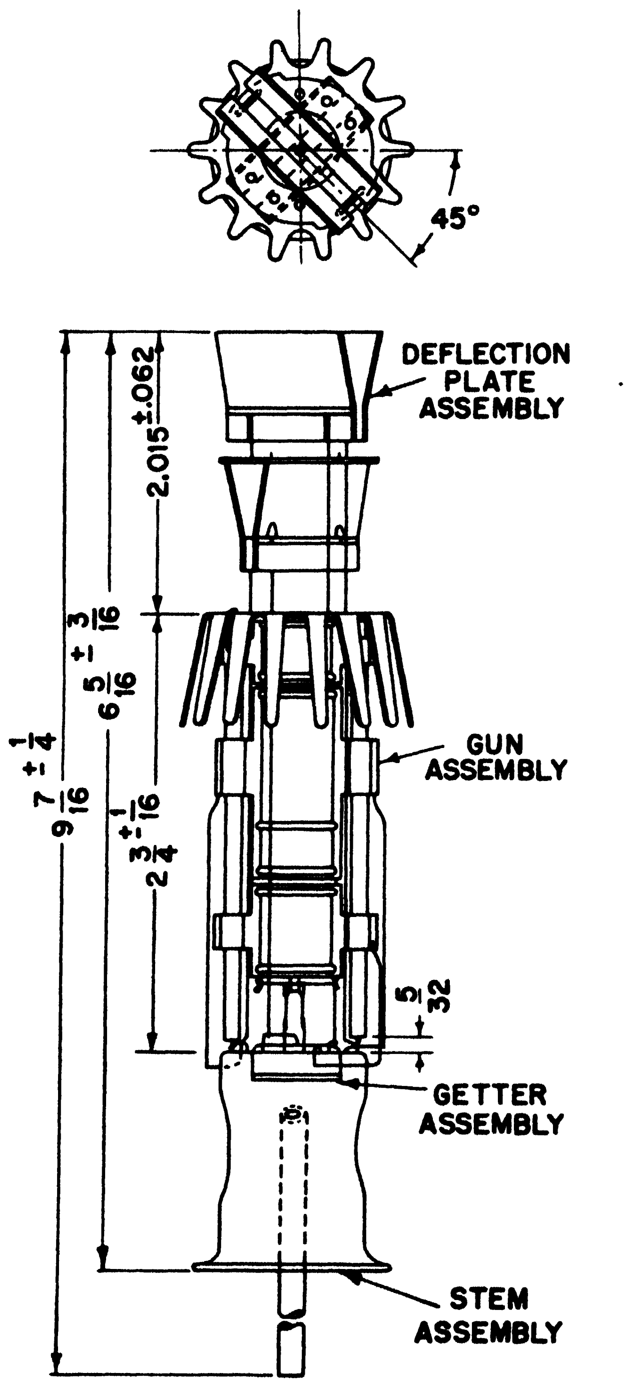
Courtesy Du Mont Labs.

Fig. 5-14.—Press seal and heater assembly of electrostatic electron gun.

are assembled on the gun structure, and are illustrated here in order to show their relative positions with respect to the gun elements. An assembly view of this typical triode gun appears in Fig. 5-13. In Fig. 5-14 is shown the stem and heater assembly. The criss-cross-wound heater fits into the opening of the cathode cylinder. A side view of the completed gun, showing the spider-shaped assembly of contacts whereby the second-anode cylinder and the coated extension of that electrode join each other, is illustrated in Fig. 5-15. This method of electrically connecting the cylinder to the painted coating extension of the anode is standard in all electrostatic tubes.

The electrical function of the different electrodes will become more understandable when we analyze the electric circuit shown in Fig. 5-16. The power-supply source for all the operating voltages is designated as *R*, this element being the voltage divider across the output of the power unit. One end of the divider is maximum negative and the other is maximum positive. The latter point is the source of the voltage applied to the accelerating anode and is *fixed* in potential; this fact should be borne in mind for future reference.

Anode 1 is connected to a point on the power supply, about one-quarter of the voltage applied to anode 2. Anode 1, therefore, is negative relative to anode 2, but positive with respect to the cathode and the control grid. Since the voltage applied to anode 1 is employed for focusing control, this voltage lead is shown affording a variable voltage. The power-supply connection to the cathode is made at some point which



Courtesy Du Mont Labs.

Fig. 5-15.—Complete triode electrostatic electron-gun assembly used in 3GP- and 5LP- cathode-ray tubes. Dimensions are in inches.

makes this electrode negative relative to the two anodes, but positive relative to the control grid. The cathode also is connected to one side or to the center tap of the heater supply.

The control grid is made negative relative to the cathode by connecting the electrode to a more negative point along the power-supply divider. A variable voltage is required at the grid since the main function is to afford control of the density of the beam, therefore, a variable tap on the divider is the source of voltage. However, the polarity of the grid must always remain negative relative to the cathode.

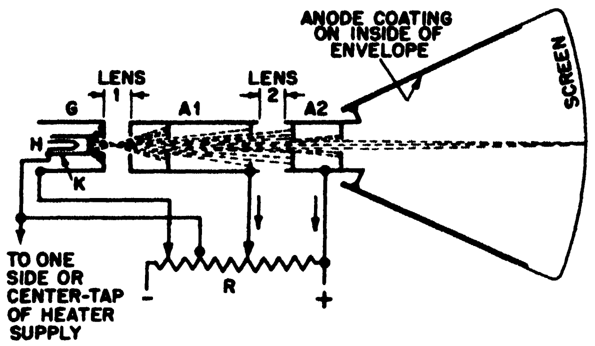

ANODE COATING ON INSIDE OF ENVELOPE
SCREEN
LENS 1
LENS 2
G
A1
A2
H
K
R
TO ONE SIDE OR CENTER-TAP OF HEATER SUPPLY
-
+

Fig. 5-16.—Schematic representation of triode electrostatic electron gun.

Examining Fig. 5-16, it is clear that the gun system is composed of four principal electrodes, exclusive of the heater; the cathode, the control grid, anode 1, and anode 2. Yet for some reason, the system is known as a triode gun. Since anode 1 also contributes to the acceleration of the emitted electrons, at least into the first lens area, the first three electrodes in the system may be the basis for the manner of typing this gun.

Analysis of the Triode Electron Gun

It is here that we refer back to the fundamentals of the triode vacuum tube. Not every action of the triode is paralleled in the electron gun, but certain principles of electrostatics are applicable to both. The matter of electrical circuitry, especially the movement of charges in the system, follows closely. In fact it might be well if we opened the discussion of the triode gun by examination of the current paths, as shown in Fig. 5-16.

Current Paths in the Triode Gun

If you examine Fig. 5-16, that is, the path of the beam as shown by the shaded areas, it is evident that not every emitted electron which finds its way through the control-grid aperture eventually reaches the screen. In fact, in most electrostatic tubes, the final beam current, or *screen current,* is only a small fraction of that which passes through the control-grid opening. No matter how well the electron lenses perform, the formation of a dense beam which will produce a bright spot of small diameter requires that stray electrons which are not tightly packed in the beam be blocked off from the deflection area and prevented from reaching the screen.

The baffles in anode 2 also contribute to restricting the boundaries of the beam so that what passes through the last aperture will be properly centered between the deflection plates. Under the circumstances, some of the electrons which are accelerated toward the screen by anodes 1 and 2 will be blocked by the baffles. Since the movement of charges between any two points is electric current, *anode current exists* between the cathode and anode 1, and between the cathode and anode 2. These are shown by the arrows adjacent to the voltage supply leads. The current flowing in the first-anode-cathode circuit is by far the greater of the two. There is no means of measuring the screen current in this case, and since the secondary emission from the screen finds its way back to anode 2, the second-anode current includes the screen current.

Action of Electrodes in the Triode Gun

The cathode K in Fig. 5-16 emits electrons due to thermal agitation of the coating material. Since this coating is located at one particular place on the surface of the cathode — facing the control-grid aperture — principal emission takes place in a forward direction from this surface. Because of the shape of the cathode, emission is not from a point source, so that while it is generally in a forward direction, it takes place in a random manner. Some of the electrons leave the cathode parallel to the axis, but off to one side. Some of them, on the other hand, leave at various angles to the axis, and still others are along the axis.

Moreover, the velocity of emission varies. Some of the electrons leave the cathode at high speed, but the majority are low-velocity charges. We might go so far as to say that the initial velocity of the emitted electron is substantially zero and any forward motion after they have left the cathode is attributable to acceleration by the voltages applied to anode 1.

So, for the present, let us act on the assumption that the positive voltage applied to anode 1 is responsible for the forward motion of the electron toward anode 1 in opposition to the negative voltage present at the control grid. Of course, electrons which are pulled into the first focusing field, and then through anode 1, are being accelerated toward the screen by the positive voltage which is applied to anode 2. This voltage not only contributes to the forward motion of the electrons, but also is actually the *main* accelerating factor. It is the voltage on anode 2 which imparts the final push to the electrons so that they arrive at the screen with sufficient velocity to excite the molecules of screen material, causing fluorescence.

Because of what will follow, it is well to stress certain pertinent conditions. The development of the first cross-over point is a function of the electrostatic field between the control grid and anode 1. Granting that this statement is broad in its implications, it is still necessary to say that any changes in potential within this field will have a material effect on the movement of charges into and through the first lens. We saw in Chapter 2 that the contour of the equipotential lines in an electrostatic field which is acting on electrons has a very great effect upon the paths taken by these electrons. Moreover, we recall in connection with the triode that both the voltage on the grid and the voltage on the plate have a great effect on the number of electrons which become the plate current.

Finally, the action in the main focusing lens formed by anodes 1 and 2 is determined by the *ratio* of the voltages applied to these electrodes; any change in either of these voltages will alter the ratio, and therefore, the focal point of this lens. In practice, focusing is accomplished by varying the first-anode voltage until the *first cross-over point is imagined on the screen.* Since the first anode also is associated with the first lens, changing its voltage cannot help but affect the two-lens systems of which it is a part. It stands to reason, then, that manipulation of the focusing adjustment is a compromise of two variables, until the best possible result is attained.

Effect of the Control Grid

Assuming for the moment that a beam of electrons exists, having a certain density and producing a spot with a certain light output on the screen, let us return to the control grid and note what effects it may have on the beam. Because of the premise that the beam exists, the voltage on the grid is some value *above* cutoff. (Referring to a *less* negative voltage in this manner is common tube practice. A voltage change which would make the control grid *more* negative than required for cutoff, would be described as a value *below* or *beyond* cutoff.)

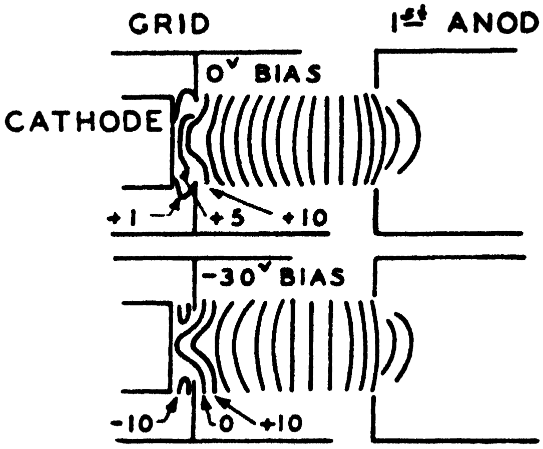
Being very close to the cathode (0.007 inch in the case of the typical triode gun in Fig. 5-12), the control grid is capable of exerting a very great influence upon the movement of charges. For example, in one case a negative bias as low as −45 volts is sufficient to offset completely the attracting influence of 375 volts on anode 1 and 1,500 volts on anode 2. Other tubes, naturally, may have different values of cutoff bias. At any rate, every cathode-ray tube has a specific grid-bias voltage which will produce beam-current cutoff for specific values of first-anode and second-anode voltages. The relationship between cutoff bias and second-anode voltage is linear in most electrostatic tubes. For any value of grid bias less negative than cutoff for the existing conditions, some beam current will flow.

In this connection, it is necessary to return to a very important point relative to conditions around the control grid. A negative voltage applied to the grid will display a great effect on the movement of electrons to and past the grid. However, this does not mean that electrons cannot reach the grid. As the consequence of the random velocity of emission, electrons can travel to the grid even in the absence of the accelerating voltages on the anodes. Electrons will reach the grid even when that electrode is made negative, thus giving rise to a slight *grid current*. This condition imposes a limitation on the resistance in the grid-to-cathode circuit because the grid current flowing through the circuit resistance will tend to increase the bias, and alter the performance of the grid electrode. Because of this circumstance, tube specifications indicate the maximum allowable circuit resistance during operation.

When tube specifications indicate maximum allowable circuit resistance, it is because of the current flow in this circuit and the voltage drop developed across the circuit resistance. When the tube is operated with a signal voltage "riding" on the grid bias, which is intended to influence the beam intensity in a specific manner, the effective grid voltage varies between some value above cutoff and zero, but it never is permitted to swing positive.

To be more specific about the action of the control grid, we might ascribe two basic functions to it. Although these were mentioned in connection with the triode tube, and find a parallel in the triode gun, they are also to be considered as applicable to every type of gun. Both are closely allied to the number of electrons which enter the first lens, and therefore, affect this formation of the first cross-over point and the density of the beam.

A change in grid voltage influences the field distribution of the first lens, and in so doing controls the emission from the cathode. For any fixed value of voltage applied to anode 1, it influences the number of electrons which pass through the cross-over point. Let us see how this comes about. In Fig. 5-17 is shown the field distribution in the first lens for two values of grid bias, 0 and −30 volts, and a fixed value of voltage on the plate.[3] It is clearly evident that with zero bias, the area adjacent to the cathode, between the cathode and the control-grid aperture, has a comparatively high positive potential as the consequence of the field between the control grid and the first anode. Under such conditions of zero grid voltage, it has been found that the area of the cathode which is emitting corresponds approximately to a projection of the area of the grid aperture; the maximum number of electrons are passing through the grid opening and the beam-current density is high.[4]



Courtesy *Rev. Mod. Phys.*

Fig. 5-17.—Field distribution in first lens for two values of grid bias.

When the control grid is made negative by an increase in the bias, −30 volts in the illustration, the field distribution in the vicinity of the cathode is altered so that only the *center* of the emitting surface is behaving as an emitter. The other areas are influenced by the space charge and effectively are not emitting. The result is a reduction in beam density and several other related effects.

Now, an unexpected condition which accompanies the increase in grid bias is a slight *increase* in the dimension of the first cross-over point over a range of bias voltages slightly above the cutoff value. This is related to the trajectories (individual paths) of the electrons emitted from the outer limits of the emitting area, and introduces another aspect which must be considered.

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[3] Morton, G. A., "Electron Guns for Television Application," *Rev. Mod. Phys.*, pp. 362-375, vol. 18, July, 1946.

[4] Moss, H., "The Electron Guns of the Cathode Ray Tube," *Jour. Brit. I.R.E.*, pp. 10-22, vol. 5, January, 1945.

If the dimensions of the cross-over point change, we can say that it is the equivalent of a change in the *sharpness* of the spot, which later is imaged on the screen by the main focusing lens. The performance of the main focusing lens is not influenced by changes in control-grid bias, and the image it will produce on the screen is essentially that which exists at the cross-over point. If for any reason, this narrowest part of the beam is not well defined, the spot on the screen, which is the image of the cross-over point, also will suffer in definition. This is similar to the photographic case, where an out-of-focus negative is enlarged; no enlarging lens can possibly restore the lost sharpness. We might add that, as the result of the change in field distribution of the first lens, due to the change in bias, the location of the cross-over point also is shifted slightly.

Since the voltages applied to anodes 1 and 2 are not affected by the grid bias, the focusing action of the main lens remains the same for the proper location of the first cross-over point, as well as for an improper location of this point. Proper focusing by a lens calls for certain specific relations between the distance of the object point and the distance of the image point, both measured from the lens. For proper relationship of the distances, the lens will focus the object to form the image. But if the image distance is held constant and the distance to the object is changed, the image no longer will be in focus.

It is because of the change in definition and the slight change in the position of the cross-over point, caused by a change in bias, that changing the "intensity" control of a triode gun system requires a slight change in the "focus" control, which varies the first-anode voltage. This action was in part responsible for the obsolescence of the triode gun and the development of other types.

Summarizing, and overlooking the undesirable effects, the action of the control grid is the control of beam density, which results in changes in the intensity of the trace on the screen. The two extremes of trace intensity are total extinction of the beam at cutoff, and the specified brilliance of the trace at those values of voltage specified by the tube manufacturer.

Changing the intensity of the trace by the application of a signal voltage to the grid is known as "intensity modulation" of the beam. The types of signals usually used for such operations resemble those which were shown as being applied to the control grid of the triode, as in Figs. 5-4 through 5-7. The shape of the intensity control signal is dependent on the manner in which the intensity is to be changed and the frequency of the signal determines the rate at which the intensity is varied.

Effect of First-Anode Voltage

Another problem encountered with the triode gun was related to the voltage applied to anode 1, the voltage being determined by the setting of the focus control. Because of the discussion of the practical electrostatic focusing system in Chapter 2, it is unnecessary at this time to dwell at great length upon the purpose of the first-anode voltage. This voltage, in conjunction with the voltage on the grid, establishes the field of the first lens; it also accelerates the charges through the control-grid aperture into the first-anode area and toward anode 2, where the second-anode voltage becomes effective. Finally, the field due to the first-anode and second-anode voltages applied to the two respective electrodes, forms the principal focusing lens.

Assuming that the voltage applied to the control grid is maintained constant, a change in first-anode voltage alters the ratio between the voltages applied to anodes 1 and 2, and changes the characteristics of the focusing lens formed between these two electrodes. Thus, varying this voltage provides focusing control. The same result could be obtained by varying the second-anode voltage, but if that were done, then both focusing and the acceleration would be changed, an undesirable condition. Moreover, since the first-anode voltage is the lower of the two, it is more convenient to work with it.

In case it is not fully understood from what was said in Chapter 2, the action of focusing by means of lens 2, controlled by the anode voltage, is a compromise function — an operation whereby the characteristics of lens 2 are modified in accordance with the changes in the first cross-over which it also affects, so that the image focal point of the lens will fall upon the screen.

Having mentioned the desired functions of anode 1 and its voltage, let us now consider several undesirable effects occasioned by variation of the first-anode voltage. These effects were among the causes of development of new types of guns, considerably modified from the triode gun. Since the first-anode voltage is common to both lens 1 and lens 2, it is understandable that any variation in that voltage, so as to effect the proper focus, would have an influence on the distribution of the field between the control grid and anode 1, and therefore, upon the characteristics of the cathode lens. This is so because the first-anode voltage affects the number of electrons which are pulled through the control-grid aperture, changing the density of the beam. Since varying the focus in the triode gun also changes the beam density, the brightness of the trace on the screen is affected. There are other effects, also, but this type of interaction between the control grid and the anode 1, as well as the converse effect of the control grid on focusing when the grid bias was varied, proved to be a nuisance, as every user of a triode-gun cathode-ray tube can well remember.

Good regulation in the power supply so as to minimize defocusing because of changes in first-anode voltage due to line-voltage loading, etc., was complicated by two related details which functioned in opposite directions. One of these was the condition that optimum focusing required a slight reduction in the first-anode voltage with increase in beam density, yet this could not be done if the regulation were sufficiently good to keep the first-anode voltage constant under varying load. Moreover, variations from tube to tube required a manual control of focusing, even for slight changes in beam intensity. The result was the development of the tetrode gun.

The Tetrode Gun

This type of gun was used during World War II and bears several different names. It is known as the tetrode, the pre-accelerator or sometimes the accelerator type, or as the screen-grid type. All, however, mean the same thing, namely that a screen-grid electrode (designated *G2* to distinguish it from *G1*, the control grid) is interposed between the control grid and anode 1, and performs several functions. Otherwise, the elements were the same as found in the original triode gun, except for changes in the physical dimensions. Although the added electrode improved the characteristics of the gun, it still was not ideal.

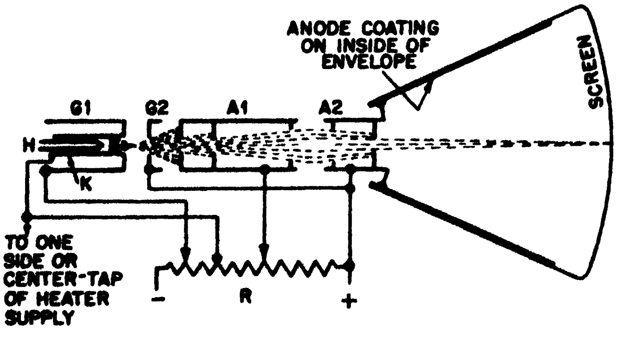
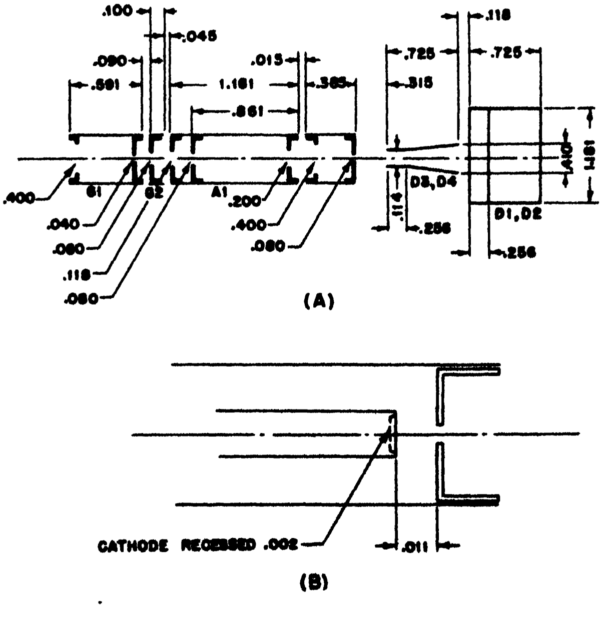

ANODE COATING ON INSIDE OF ENVELOPE
SCREEN
G1
G2
A1
A2
H
K
TO ONE SIDE OR CENTER-TAP OF HEATER SUPPLY
-
R
+

Fig. 5-18.—Schematic representation of tetrode electrostatic electron gun.

The electric circuit of the tetrode gun is shown in Fig. 5-18. The screen grid, or pre-accelerator electrode, *G2* is directly connected to anode 2, resulting in several effects. First, the higher potential of the electrode following the control grid accelerates the electrons to a potential equal to that of the second anode. This makes for a denser beam, since more current is drawn from the cathode at any rated value of control-grid voltage than in the case of the triode, for like values of second-anode grid voltage. Another result is a smaller diameter of the beam, because of a smaller first cross-over. Thus, a smaller and brighter spot is produced on the screen of the tube.

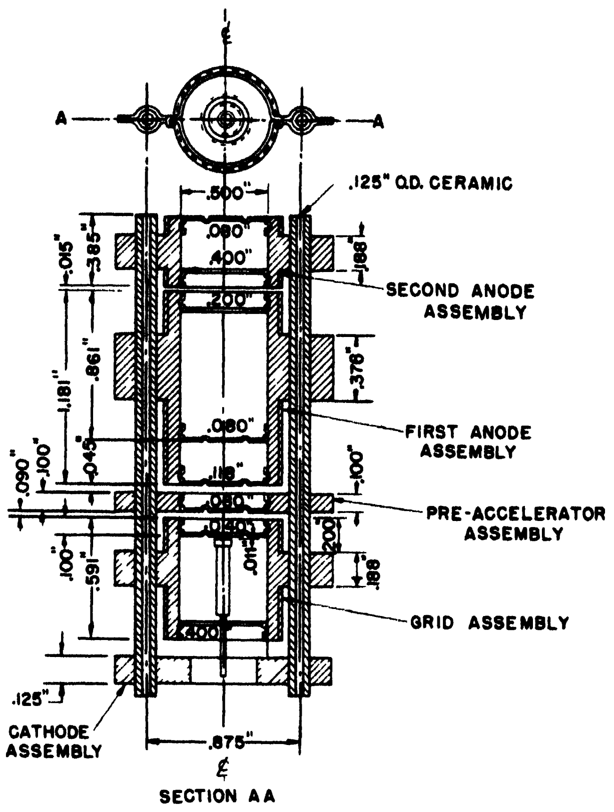
Still another effect is the elimination of interaction between the control grid and the first anode. Changes in the potential of the focusing anode do not influence the field responsible for the first cross-over, since now this field exists between the control grid and the pre-accelerator grid, which is at a constant potential. Therefore, changes in the focusing voltage do not change the beam density; all that is affected is the ratio of voltages applied to anodes 1 and 2.

A detail drawing of a typical tetrode gun is shown in Fig. 5-19, in which the screen grid can be seen to be a very shallow disk. The constructional features of the gun and the mounting of the electrodes are more clearly illustrated in Fig. 5-20.



Courtesy Du Mont Labs.

Fig. 5-19.—Dimensional drawing of tetrode electrostatic electron gun used in 3BP- and 3FP- cathode-ray tubes. Detail of cathode-to-grid spacing is shown in (B). Dimensions are in inches.



Courtesy Du Mont Labs.

Fig. 5-20.—Sectional view of tetrode electrostatic electron gun.

Although an improvement over the triode, the pre-accelerator type of electrostatic gun is not the final development. It was modified because of the current drain of the first, or focusing, anode. Although the effects of changes in first-anode voltage are less serious in the tetrode type of gun, it is nevertheless necessary to have good regulation in the power supply. In fact, the very existence of current drain by the focusing anode resulted in a waste of power, which was aggravated in this design, as in the triode, by the need for a bleeder of comparatively low resistance across the power supply. Eventually, there was a change in the design of the gun, which produced the zero-first-anode-current type.

The Zero-First-Anode-Current Gun

The main advantage of this type of gun is, as its name implies, the elimination of any current drain by the focusing anode. The gun was first introduced by RCA about 1937; its use in certain cathode-ray tubes which originally used a triode or tetrode gun, is indicated by the presence of a suffix letter "A" after the regular type number. Schematically, the gun is as shown in Fig. 5-21. The electrical connections are the same as in the screen-grid type.

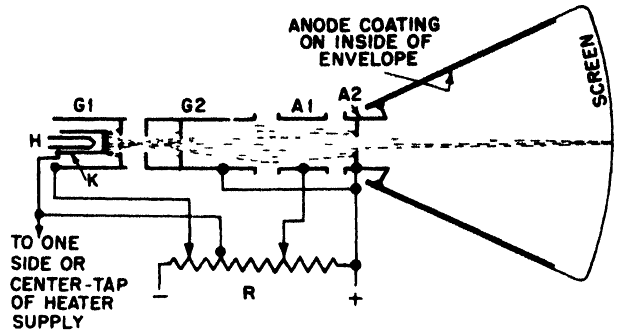
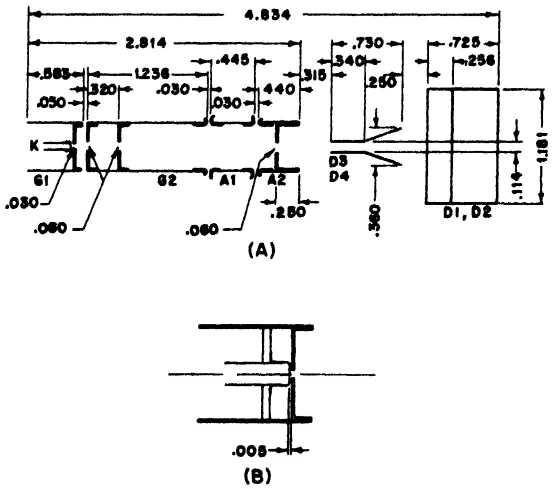

ANODE COATING ON INSIDE OF ENVELOPE
SCREEN
G1
G2
A1
A2
H
K
TO ONE SIDE OR CENTER-TAP OF HEATER SUPPLY
R
−
+

Fig 5-21.—Schematic representation of zero-first-anode-current electrostatic electron gun.

Fig. 5-22 shows the physical construction of the gun; it can be seen that the spacing between the cathode *K* and the control grid *G1* was materially reduced. The pre-accelerator grid *G2* was changed from a shallow to a long cylinder containing a masking baffle, and the original long-cylinder focusing anode *A1* was modified to a comparatively short cylinder without any baffles. The accelerating anode *A2* was changed to a cylinder containing a single aperture baffle.

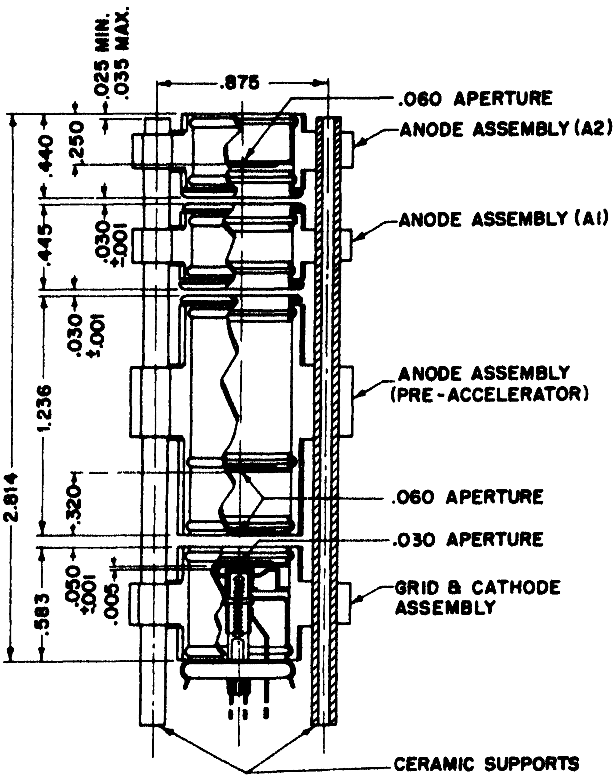
Relative to the beam, several advantages are gained. First the beam is made narrower before it enters the focusing area. This can be noted by comparing the electron paths in Figs. 5-18 and 5-21. Second, the stray electrons in the beam, which might tend to



Courtesy Du Mont Labs.

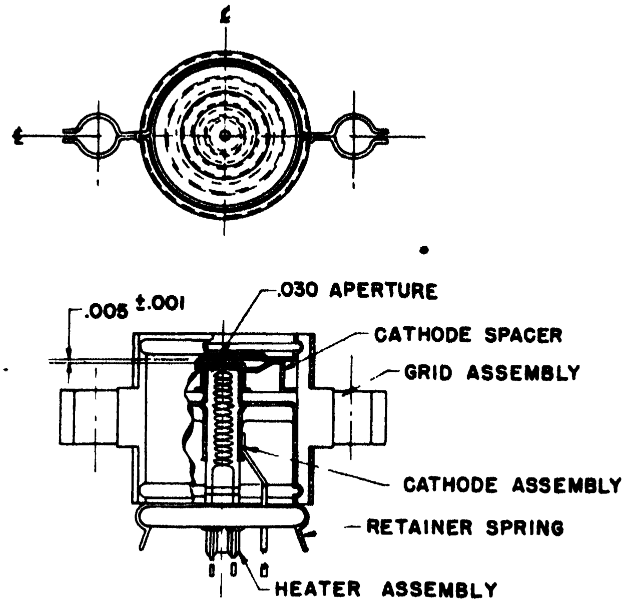
Fig. 5-22.—Dimensional drawing of zero-first-anode-current electrostatic electron gun used in 5CP-A cathode-ray tube. Enlarged view of cathode-to-grid spacing is shown in (B). Dimensions are in inches.

broaden the spot, are prevented from entering the second-anode area and passing through its aperture. Third, since the focusing anode contains no baffles and the limits of the beam are fairly well confined while passing through that electrode, the surface of the focusing anode does not intercept any electrons, hence draws no current.



Courtesy Du Mont Labs

Fig. 5-23.—Sectional view of zero-first-anode-current electrostatic electron gun. Dimensions are in inches.



Courtesy Du Mont Labs.

Fig. 5-24.—Sectional view of grid and cathode assembly of zero-first-anode-current electrostatic electron gun used in 5BP-A cathode-ray tube. Dimensions are in inches.

Since the design features of the gun in the neighborhood of the control grid are like those of the pre-accelerator type, freedom from interaction between the control grid and the focusing anode exists. The assembly of a typical zero-first-anode-current gun is illustrated in Fig. 5-23, and an enlarged view of the cathode assembly showing the arrangement of the heater is illustrated in Fig. 5-24. Except for the dimensions, the relative positions of the heater element and the remainder of the cathode assembly apply to the previous types of electrostatic gun assemblies.

Concerning the constructional illustrations of electrostatic guns as shown herein, it must be noted that the dimensions apply to the product of the specific manufacturer indicated. Other manufacturers' products may perform in a standard manner, but the actual constructional details for similar types of tubes may be somewhat different. We have attempted to show only typical examples of such details.

Fixed-Focus Gun

A modification of the fundamental triode gun which more readily keeps the focus fixed, or at least minimizes the number of variables which will affect the focusing of the beam, is described by Iams[5] and is

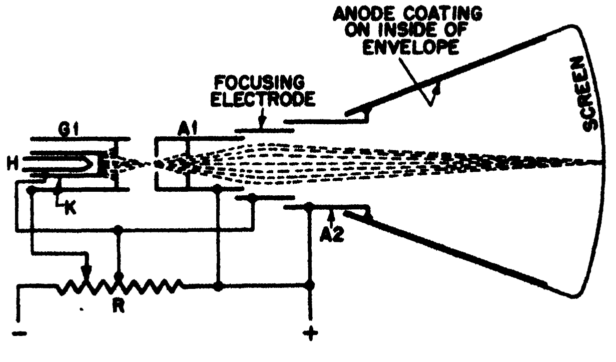


Fig. 5-25.—Schematic representation of fixed-focus electrostatic electron gun.

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[5] Iams, H., "A Fixed-Focus Electron Gun for Cathode-Ray Tubes," *Proc. I.R.E.*, pp. 103-105, vol. 27, February, 1939.

shown in Fig. 5-25. The focus is made independent of anode voltage by using, for the electron lens, electrodes at cathode potential and at a common anode potential. In this case, the electrons from the cathode are controlled and accelerated into the anode region just as in the original triode gun shown in Fig. 5-16. The combination of electrostatic fields between the focusing electrode and the sections of the anode (anode 1 and 2) gives to the outermost electrons of the bundle a greater velocity toward the tube axis than is given to the electrons within the bundle. When the length of the focusing electrode is great, or its diameter small, the beam is brought to a focus at any desired distance from the gun.

Once the electron beam has been focused in this way, it remains focused regardless of the anode voltage applied. This fact may be understood by regarding the electrostatic focusing field as a means of deflecting the electrons in the beam so that all meet at one point. In the fixed focus electron gun, the focusing field occurs between electrodes at cathode and anode potentials. For a given angle of deflection, the field required is directly proportional to the voltage through which the electrons have been accelerated. The field deflecting the electrons to the point of focus increases linearly with the anode voltage. Hence, the path of an electron moving under the influence of the focusing fields in the electron gun is unique regardless of the anode voltage.

Intensifier Anode

On several occasions, reference has been made to the intensifier or post-deflection acceleration anode. Although not a part of the gun and more intimately associated with the deflection system than with the gun, it does have a logical connection with the second anode, especially in connection with the ratio between the voltages of anode 2 and the intensifier ring.

One of the problems encountered in the operation of the cathode-ray tube is that of obtaining sufficient trace brilliancy. Certain uses demand an exceedingly bright trace for daylight viewing or for projection purposes. This is not limited to television applications but often is a requirement in laboratory or classroom demonstration equipment.

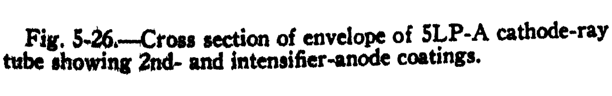
ELECTRON GUN AREA
DEFLECTION AREA
A2
SCREEN
2ND ANODE COATING ON INSIDE OF ENVELOPE
INTENSIFIER ANODE COATING ON INSIDE OF ENVELOPE
INTENSIFIER CONNECTOR

Fig. 5-26.—Cross section of envelope of 5LP-A cathode-ray tube showing 2nd- and intensifier-anode coatings.

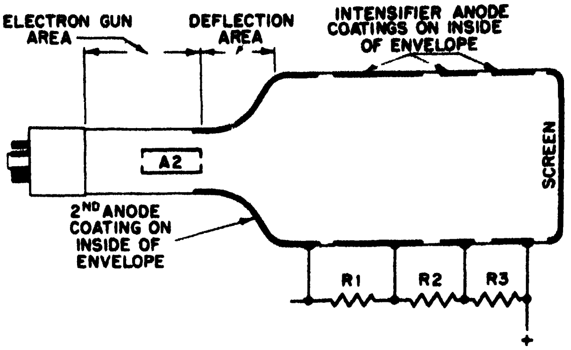
ELECTRON GUN AREA
DEFLECTION AREA
INTENSIFIER ANODE COATINGS ON INSIDE OF ENVELOPE
A2
SCREEN
2ND ANODE COATING ON INSIDE OF ENVELOPE
R1
R2
R3
+

Fig. 5-27.—Cross section of envelope of 5RP-A cathode-ray tube showing 2nd- and intensifier-anode coatings. Relative potentials are determined by a voltage divider consisting of resistors *R1*, *R2*, and *R3*, all of which are equal.

The determination of beam velocity is the function of the second-anode voltage; the higher this potential, the higher is the energy conveyed by the moving charges, and consequently, the higher the brilliancy of the trace on the screen of the tube. However, there is one disadvantage. The location of the second anode is ahead of the deflection system in the tube, as indicated in Figs. 5-26 and 5-27. If the electrons in the beam are given sufficient velocity by the second-anode voltage to strike the screen with the necessary force to produce a very bright trace, they travel through the deflection area at a very high velocity. Since deflection voltages must increase as the electron velocity becomes higher, the development of the deflection amplifiers may present undue complications. The use of a high-intensity beam restricts the usefulness of the tube, because the sensitivity to deflection voltages is materially reduced. A means for producing the required trace intensity without decreasing the deflection sensitivity is by the use of post-deflection acceleration. In this arrangement, the velocity of the beam is increased by the application of a high voltage E_{bs} to the intensifier anode, *after* the electrons have been deflected. In contrast, the normal condition in most electrostatic cathode-ray tubes is high acceleration of the beam prior to deflection.

The above statement may seem to imply that the tube having an intensifier anode, for post-deflection acceleration, does not contain the conventional second anode.

The usual second anode is used, however, except that the voltage applied to it is not as high as would otherwise be required to produce the required trace intensity; in that way, the deflection sensitivity is kept as high as possible. For providing additional acceleration to the electrons, after deflection, the intensifier anode is used; it is located in the region between the deflection plates and the screen of the tube.

It might appear from the above that post-deflection acceleration affords only advantages, thereby making the tube which does not utilize this system inferior to one that does. Such is not the case. The intensifier anode requires a higher voltage source than would otherwise be necessary, or perhaps a separate power supply. This adds to the cost, but if the need for sensitivity and trace brilliance cannot be satisfied with the conventional tube, then post-deflection acceleration becomes necessary.

The other side of the story is that the use of post-deflection acceleration tends to cause a slight amount of distortion in the spot, which is, however, kept to an acceptable minimum by keeping the ratio of $E_{b3}/E_{b2} = 2$ or thereabouts. Exceeding this limit in tubes not intended for such operation gives rise to undesirable effects, described in the section of this chapter which deals with operating voltages.

These post-deflection accelerating anodes are rings or wide bands of electrically conductive material painted on the inside surface of the envelope. They are insulated by a gap from the extension of the second anode, which is similarly applied to the inner surface of the envelope, extending as a rule from the location of the second anode in the gun, almost to the screen of the tube. When the tube utilizes but one intensifier ring, it is located between the deflection area and the screen, nearer to the screen. In those cases where several such rings are used (Fig. 5-27), they are concentrically arranged on the inside wall, each ring insulated from the other and each being at a progressively higher voltage. These rings are insulated from the coated extension of the second anode which, in this case, would terminate below the first intensifier anode coating, but each ring is joined to the other by a very high resistance coating inside of the tube. The distribution of progressively higher intensifier anode voltages to the different anodes is made as shown in Fig. 5-27.

Due to its location, the intensifier anode (or anodes), receives the secondary emission from the screen; since it does not attract electrons from the high velocity beam which passes it, the current in the intensifier-to-cathode circuit is considered to be the screen current. Connection to the intensifier ring or rings is made by way of connecting terminals which protrude through the envelope.

ELECTRON GUNS IN ELECTROMAGNETIC TUBES

Electron guns used in magnetic-type tubes are similar in many respects to the electrostatic variety, yet there are, naturally, points of dissimilarity. Like the electrostatic type, the magnetic tube gun requires an electron beam for its operation and since the production of the display — the excitation of the screen by an electron stream — is the same in both, it is not surprising that certain of the control actions are very much alike in both types of guns. Both types develop the first cross-over in the same fashion, by means of an electrostatic field between the control grid and an adjacent accelerating or shielding electrode, according to the type. Beyond that point, the magnetic tube gun differs radically.

The differences can be divided, for purposes of discussion, into two main groups. One is operational and the other is physical. The operational differences embrace the means used for the main focusing, and the separation of ion flow from electron flow. The physical differences are related to the actual mechanical construction of the gun assembly, and are especially important in considering those which are of special construction.

Virtually all magnetic tube guns make use of both magnetic focusing and deflection. At the time of this writing, only a few tubes employ electrostatic focusing and magnetic deflection. Although this arrangement should logically fall in the category of the electrostatic type of gun previously discussed, it has enough singular features to warrant a separate treatment, which will be given after the discussion of the magnetic tube gun.

Like the electrostatic gun, the magnetic tube gun assembly has passed through a series of improvements, necessitated by changes in requirements, such as tube dimensions, changes in screen types and others. In spite of the changes, the general principles of the gun remained substantially the same. Some of the changes in magnetic-gun types were founded on the same needs as were described in connection with the electrostatic types, especially the interaction between functions of the control grid and the accelerating anode.

The parallel between the magnetic- and the electrostatic-type guns in many respects is so close that previous comments concerning the triode vacuum tube will be found to be applicable.

The Fundamental Magnetic Tube Gun

A representation of the fundamental type of electron gun used in a cathode-ray tube which employs magnetic focusing and magnetic deflection is found in Fig. 5-28. The assembly of the heater *H*, the cathode *K*, the control grid *G*, and the accelerating anode *A*, is like that used in the electrostatic gun, hence requires no discussion concerning the arrangement. There are, however, variations, and not every magnetic-type tube has the accelerating anode in this position.

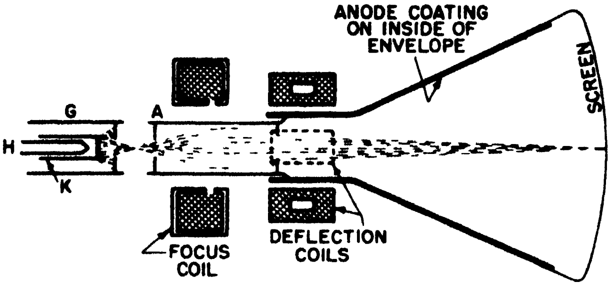

Fig. 5-28.—Fundamental electromagnetic electron gun.

At any rate, the accelerating anode *A*, which is a part of the electron gun, is electrically connected to an extension of that electrode, which is in the form of an electrically conductive coating painted on the inside wall of the tube cone. Connection between the gun anode and the coated anode is made by means of a "spider" on the gun mount, just as in the electrostatic variety. Around the outside of the tube envelope and in the proximity of the anode barrel is the magnetic focusing coil. This structure is responsible for the principal focusing of the stream of electrons.

The focusing coil is not a part of the actual gun assembly, being outside of the tube envelope; yet it still is necessary, in discussion, to consider it as being part of the gun, because focusing of the beam is one of the functions of the gun. The illustration also shows the deflection coils, which will be discussed in the next chapter.

The general pattern of the electron beam path in such a gun is shown in Fig. 5-28. As can be easily seen, it is not widely different from the pattern associated with the electrostatic type of gun. This is to be expected, since the origin of the beam and the end result are the same. A few minor variations will appear, as different types of magnetic tube guns are described, but in general, all resemble that which is shown in the figure just mentioned. In all cases, the spot on the screen is the image of the first cross-over with the possible qualification that the image may be smaller than the first cross-over because of the action of the limiting apertures in the accelerating anode barrel.

The similarity in the organization of the fundamental magnetically focused gun to the conventional triode tube is striking. So alike are they that it is not necessary to comment at great length concerning the relative actions of the gun elements. About the cathode-control-grid relationship we need say very little; here too, as before, the control grid controls electron emission and the intensity of the beam between the limits of zero beam current and the value giving full rated trace intensity or light output at the screen. The control grid determines the number of electrons which pass through its aperture, under the influence of the voltage applied to the accelerating anode, to form the first cross-over point.

The function of the accelerating anode likewise requires little comment. With a positive voltage applied, it accelerates the electrons from the first lens area, through the control-grid aperture, into the main focusing area and thence through the deflection area, to the screen of the tube. With an accelerating voltage substantially higher than is usually found in electrostatic-type tubes, several very desirable effects are attained.

Advantages of Magnetic Tube Guns

The first advantage of guns using magnetic focusing and deflection is increased rate of electron flow in the beam at the first cross-over point giving a very much brighter trace. The second effect is a reduction in the diameter of the cross-over in the cathode lens, thus attaining higher luminescence and smaller spot size. These very important advantages are characteristic of magnetic-type tubes, because the system permits the application of a high accelerating voltage in the neighborhood of the cathode; for this reason, applications which demand maximum resolution in the display, such as television and the many systems which make use of radar principles, utilize the magnetic type of tube to full advantage.

Projection television and other systems require higher than usual brightness at the screen. Since the dimensions of the optical projection system should be as small as possible, to minimize the cost, it is advantageous to use a cathode-ray-tube screen of small diameter. In order that proper resolution be available with such a small-sized screen, it is imperative that the spot diameter be small. For example, in the 3NP4 tube used in Protelgram television projection system, the picture size on the projection tube is only 1.42 ×

1.89 inches for 525 scanning lines. The maximum tolerable spot diameter, therefore, is 1.42/525 or 0.0068 cm.[6]

Generally speaking, spot dimensions increase with screen dimensions, even though identical guns may be used in tubes of different sizes. It is a matter of the various voltages which are applied to the different tubes which is responsible for the change in the cathode lens behavior, the first cross-over dimension, and finally the image on the screen.

Magnetic Focusing

The basic principles of the focusing action were described fully in Chapter 3, but will be reviewed briefly here. The main focusing action, caused by the magnetic field issuing from the focusing coil, is the application of such rotational forces to the electrons in the beam, as to make those which left the first cross-over at an angle to the axis return to the axis at some point which represents the focal point of the magnetic lens. These rotational forces are proportional to the angle of divergence from the axis at which the electrons entered the field. In this way, theoretically, all the electrons which entered the field, after having left a point source located on the axis, return to the axis. The point where they return when properly focused under the usual magnetic lens conditions, is upon the screen.

Focus Coil

Fig. 5-29 shows a typical commercial focus coil. Although there are various types and sizes of coils to suit individual needs, the sample shown will serve adequately for discussion. As briefly described in Chapter 3, the focus coil is a multilayer polenoid; a great many turns are wound on a form of short length to provide an approximately uniform field within the core. The field of such a coil, in cross-section, is shown in Fig. 5-30A. If, however, the coil is encased in a magnetic material except for the inner surface, as in Fig. 5-30B, the effective length of the field is shortened. A greater concentration of the magnetic field is obtained when the coil is totally enclosed except for a small gap. This field distribution is shown in Fig. 5-30C, and is typical of most commercial coils. Focus coils producing different field configurations are manufactured for special applications, projection television, etc. Although most of these merely utilize variations in the magnetic case, certain types are available with adjustable air gaps rather than the fixed gap illustrated.

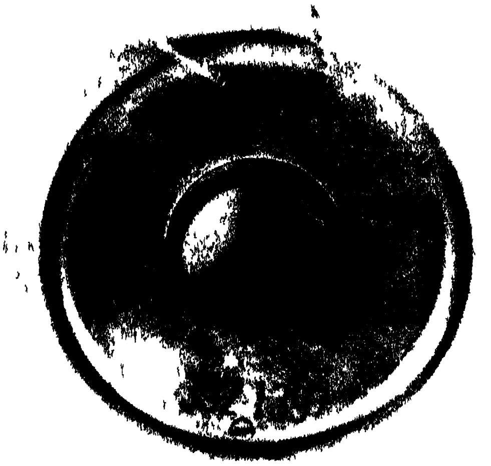

Fig. 5-29.—Typical focus coil for electromagnetic cathode-ray tube.

Unlike the electrostatic-focusing system wherein the main focusing cylinder assembly is fixed in position at

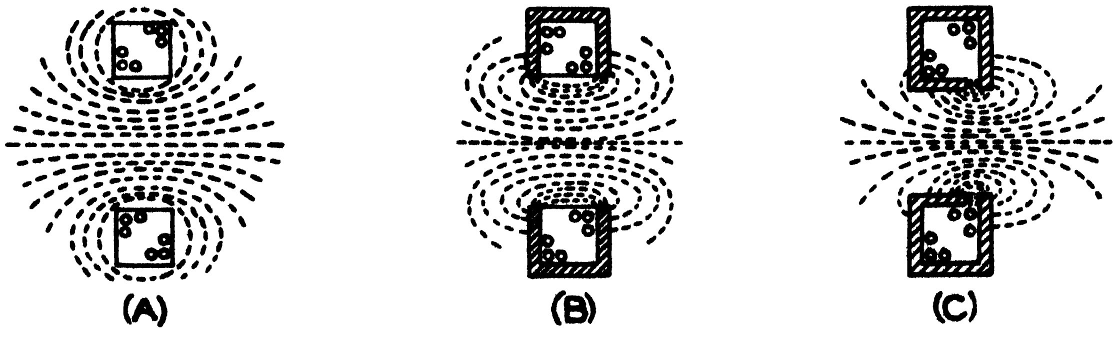

Fig. 5-30.—Magnetic fields within: (A) unshielded focus coil; (B) focus coil surrounded, except for inner surface, by magnetic shield; (C) focus coil surrounded, except for small gap on inner surface, by magnetic shield.

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[6] Rinia, H., de Gier, J., and van Alphen, P. M., "Home Projection — Part I. Cathode-Ray-Tube and Optical System," *Proc. I.R.E.*, p. 395, vol. 36, March, 1948.

the time the gun is manufactured, the magnetic focus coil is placed over the neck of the tube at the time of installation and then adjusted for best performance.

Effects of Focus-Coil Position

The positioning of the focus coil controls a number of operational details. The primary one is the relationship of the object and image distances for the stated operating conditions, so as to bring the electrons from the object (first cross-over) to a focus on the screen. From this point of view, the exact location is not highly critical, since the adjustment range of focusing current will allow shortening or lengthening of focal distances so as to compensate for other influences.

The positioning of the focus coil also affects to some extent the spot size; the optimum location is that which produces the smallest spot consistent with the required brilliancy, sharpness and roundness, and which permits control of focusing current on both sides of the desired spot condition. Moving the coil away from the optimum location tends to increase the spot size.

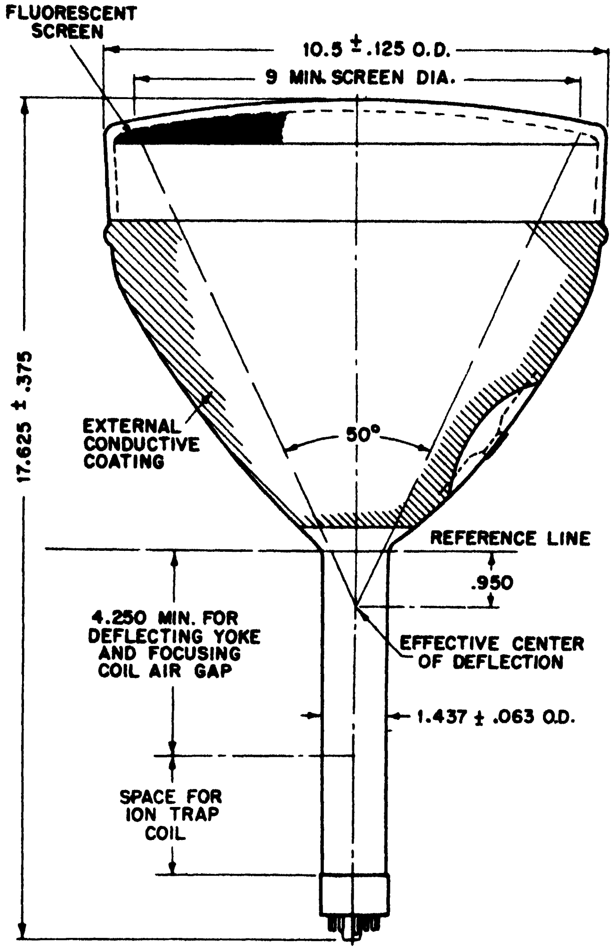
Another detail associated with positioning of the focus coil is its effect on the sharpness of the spot as it is deflected from the center toward the useful limits of the screen. Certain positions of the coil may afford the desired spot condition at the center of the screen, but there is a loss of sharpness as the spot moves from the center. Naturally, the proper positioning is that which affords the best spot condition across the entire useful surface of the screen.

The position of the focus coil also is related to the correction of misalignment of the gun electrodes, which results in an undeflected and unfocused spot at some point on the screen which, while within specified limits, is not at the geometric center of the screen. The misalignment will vary from tube to tube, between such extremes as from perhaps a circle with ½-inch radius to one which has a 1½-inch radius. The tolerance in misalignment usually is stated by the manufacturer in the tube specifications. The correction of the centering by means of the focus-coil position frequently requires slight tilting of the focus coil. Moving the coil forward and backward along the length of the tube neck will affect the spot, but usually will not correct appreciable misalignment. Since most focus-coil assemblies are mounted upon yoke supports, tilting the plane of the focus coil slightly, so as to center the beam properly, can be done quite conveniently.

Tube Reference Line

For approximate positioning of the focus coil, tube charts or specifications usually are accompanied by an outline drawing of the tube giving various dimensions. Among these dimensions is the location of the "reference line," a term which generally appears in the description of the focus-coil positioning.

An example of such a drawing is shown in Fig. 5-31.



Courtesy North American Philips

Fig. 5-31.—Outline of 10BP4 electromagnetic cathode-ray tube.

This is an outline drawing of a magnetic-type tube, the 10BP4, from which many details not pertinent to the present discussion have been deleted. The reference line is shown and is that position where a cylinder two inches long and of appropriate inside diameter will come to rest on the *cone* after being slid *over the tube neck* as far as it will go. For tubes having a nominal outside neck diameter of 1⅜ (1.375) inches, the cylindrical standard has an inside diameter of 1.430 (+0.003, −0.000) inches. For tubes with a nominal neck diameter of 1⁷⁄₁₆ (1.4375) inches, as the 10BP4, the inside diameter of the test cylinder is 1.500 (+0.003, −0.000) inches.

Focusing Control

The use of current as a means of focus control introduces a new quantity to work with. Heretofore, in electrostatic systems, we worked with values of voltage as a means of varying the focus. In the magnetic system, the control is accomplished by changes in current, which in turn vary the intensity of a magnetic field. Although it is understandable that current will not pass through a coil unless a voltage is applied across its terminals, it is common practice to speak in terms of current, rather than voltage, when referring to the action of the focusing coil. It is assumed that the necessary d-c voltage source to produce the desired current through the coil exists, or that the coil will be connected in series with a circuit carrying the required amount of direct current

Two methods of specifying the focus requirements for magnetically focused cathode-ray tubes are commonly used. One of these states the amount of *current* which must pass through the focus coil in order to focus the beam under a given set of operating conditions. This current applies, however, only if the standard type of focus coil, specified with the tube data, is used.

If a focus coil other than the standard type is to be used, then the *ampere-turns* rating of the coil must be considered. This specification is much more general in its application because it affords a latitude in design for any one system. H, the intensity of a magnetic field (number of lines of force per unit area) produced by a current-carrying coil is equal to $0.4\pi NI$. From this, it can be seen that the magnetic force developed by the coil is proportional to the number of turns, N and the current, I. The product of the two, $N \times I$, equals the ampere turns. Thus, a coil of 4,000 turns carrying 0.15 ampere is rated at 600 ampere turns. The same magnetic field can be produced with a coil of 2,000 turns and 0.3 ampere, since the resultant product is again 600 ampere turns.

This approach makes it possible for the coil designer to select such coil construction specifications as best suit his purpose. For example, the specification for a 10BP4 tube calls for a focus-coil current of 135 milliamperes for a certain grid voltage and an accelerating voltage to produce a specified light output on a six-by-eight-inch raster on the face of the tube, using a standard focus coil. For any other focus coil the specification is 4,500 ampere turns, with the same tube voltage and light output ratings.

Regardless of what method of focus-coil rating is used, an appropriate source of voltage is needed. The value of this voltage is, of course, determined by the amount of current required and the resistance of the coil through which it must flow.

It is almost meaningless to quote average ampere-turn ratings of focus coils used with present-day tubes, except perhaps as it may apply to quite a few tubes which were used during World War II in different radar applications. The majority of these operated with focus coils approximating 400 ampere turns, but the television tubes in use today require about 550 to 600 ampere turns. An exception is the 10BP4, for which the figure of 4,500 ampere turns was stated previously.

Relationship Between Focus-Coil Requirements and Accelerating Voltage

Although the mathematical analysis of the relationship is beyond the scope of this text, it can be shown that the required magnetic focusing field is proportional to the square root of the accelerating voltage. Since the field intensity is proportional to the ampere turns of the coil producing the field, the ampere turns will then be proportional to the square root of the accelerating voltage. However, for small variations of accelerating voltage, the change in the ampere-turns rating is proportional to one-half the change in accelerating potential.[7] In order to vary the ampere turns of the focus coil, either the current or the number of turns may be changed. Under normal conditions, it is the focus-coil current which is varied.

Commercial Examples of the Fundamental Triode

The fundamental triode gun is used in two projection-type television tubes now on the market. One of

Courtesy National Union

Fig. 5-32.—TP400A television projection tube.

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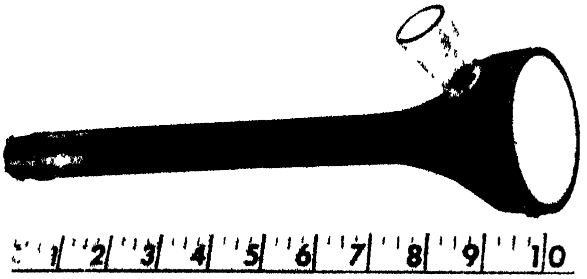
[7] Soller, T., Starr, M. A., and Valley, G. E., Jr., "Cathode Ray Tube Displays," 1st ed., pp. 93-105, McGraw-Hill Book Co., Inc., New York, N. Y., 1948.

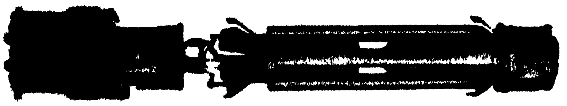
these, shown in Fig. 5-32, is the TP400A, a 4-inch projection tube, designed for an accelerating voltage of 20 kilovolts. The complete specifications will be found in the Appendix.



Courtesy North American Philips

Fig. 5-33.—3NP4 television projection tube.

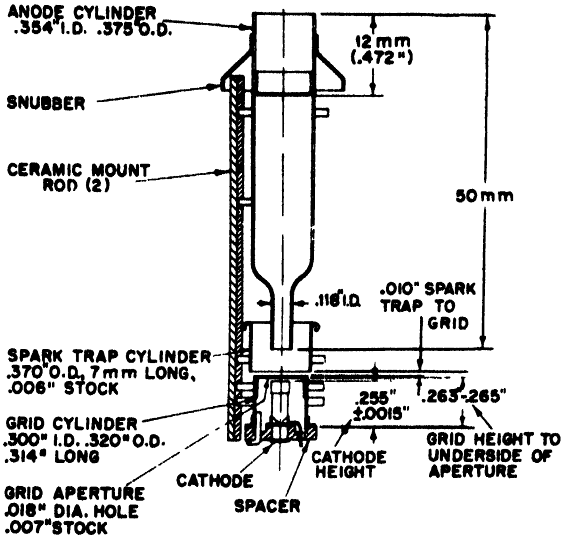
The second example, shown in Fig. 5-33, is the 3NP4, also used in projection-type television receivers. This tube has a 2½-inch diameter face and operates with a potential of 25 kilovolts on the accelerating anode. The glass cup on the cone of the tube houses the high-voltage anode button, and serves to lengthen the external leakage path from the high-voltage contact to the deflection and focusing coils. While discussing this tube, we might mention the external conductive coating present on this and many other magnetic tubes, although it is not related to the electron gun. This coating is usually grounded. Since the second-anode coating is applied to the inner surface of the glass envelope, the two coatings form a capacitor, which is connected across the output of the high-voltage supply and functions as a supplementary filter capacitor. Reference was made to this in Chapter 4 as a subject among the mechanical characteristics.



Courtesy North American Philips

Fig. 5-34.—Electron gun used in 3NP4 cathode-ray tube.

Fig. 5-34 is a photograph of the electron gun used in the 3NP4, while a diagram of this assembly appears in Fig. 5-35. The cathode and control grid are conventional in shape. However, the conformation of the anode cylinder is substantially different from the usual anode found in other type tubes. Because of the unusually small dimensions of the tube, and the high voltages involved, an electrode identified as a "spark-trap cylinder" is positioned between the anode and the control grid. The narrow neck of the anode projects into this cylinder, which is grounded. The function of the spark trap is to protect the grid circuit, should an intermittent arc jump the gap between the anode and any adjacent point. Since the spark trap is closer to the anode than any other element, it will attract any spark which may occur from the anode cylinder.



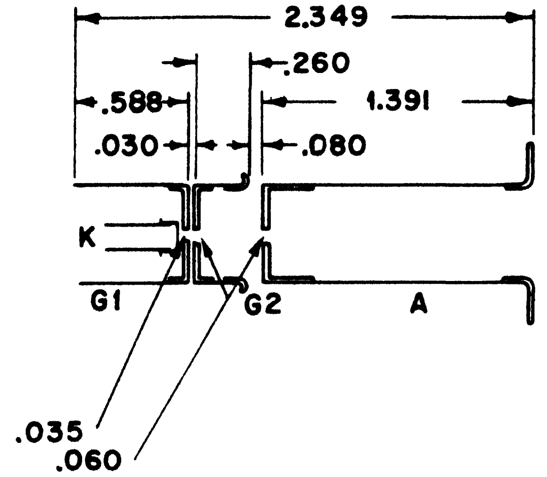
Courtesy North American Philips

Fig. 5-35.—Sectional view of triode electromagnetic electron gun used in 3NP4 cathode-ray tube.

The Tetrode (Type I)

If, to the basic triode just discussed, a second grid is added, the tetrode electron gun is achieved. Despite the fact that there are four elements, this gun is sometimes referred to as a "triode" type; this text will consider it as a tetrode. The reference to type I in the section heading applies to the construction of the accelerating anode cylinder.

The organization of the gun, showing details of dimensions and spacing, is illustrated in Fig. 5-36. The cathode *K* and control grid *G1* have a configuration along the general lines shown previously. The new element, screen grid *G2*, is a relatively shallow cylinder, having an aperture at the end facing the control grid and completely open at the end which faces the accelerating anode *A*. Because of the shape of the anode, this gun was identified as "nonlimiting aperture" type and was commonly used in radar equipment employed during the last war.

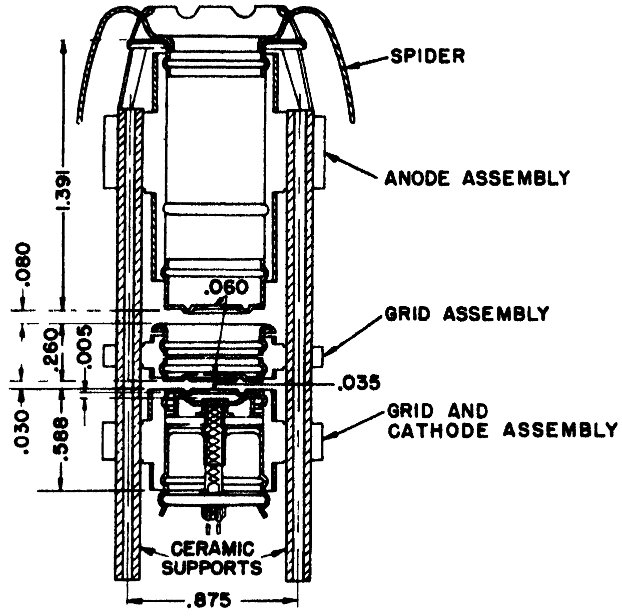


Courtesy Du Mont Labs

Fig. 5-36.—Dimensional drawing of tetrode electromagnetic electron gun. Dimensions are in inches.

The function of the screen grid is twofold. First, it isolates the control grid from the accelerating anode, preventing interaction between the two, and so making the control action of *G1* virtually independent of the voltage applied to *A*. Second, the first cross-over point is developed as the consequence of the field between the control grid and the screen grid, the latter being at approximately 250 volts positive, the exact value being dependent upon the specific tube. Consequently, changes in anode voltage have very little effect on the cutoff bias or other actions of the control grid. However, changes in the voltage applied to *G2* have sufficient influence on the beam, so that this electrode could be, and is, used for beam control in addition to the control grid

While the screen isolates the control grid from the anode, the relative positions of this element and the anode have a bearing on the dimensions of the beam which traverses the gun and advances to the screen. The higher the accelerating voltage relative to the voltage on the screen grid, the narrower the cone of electrons which advances through the gun and the smaller the dimensions of the spot on the screen. Since the accelerating voltage also determines beam velocity, increasing this voltage within operating limits increases the brilliance of the trace. With the accelerating voltage and the beam current constant, an increase in voltage applied to *G2* will also reduce the spot dimensions.



Courtesy Du Mont Labs

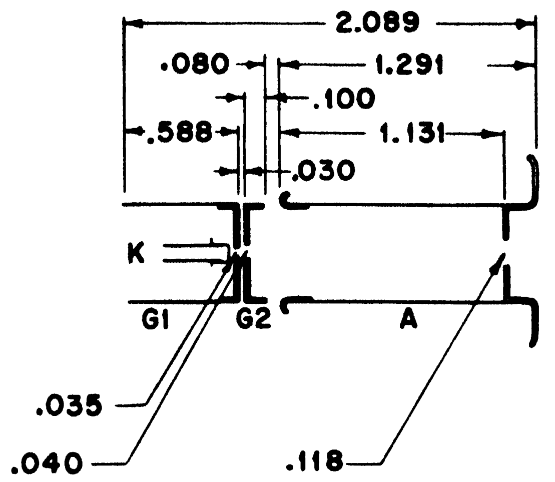
Fig. 5-37.—Sectional view of tetrode electromagnetic electron gun used in 5FP-, 7BP-, and 12DP- cathode-ray tubes. Dimensions are in inches.

An assembly drawing of the tetrode gun is shown in Fig. 5-37. The general construction is very much like that used in the electrostatic type; in this figure, the location of the heater element within the cathode sleeve can clearly be seen. The spider used for making contact between the anode electrode in the gun and the anode coating on the inside of the tube envelope is seen joined to the anode barrel at the top of the figure.

Standard Television Tetrode (Type II)

A modification of the tetrode gun is illustrated in Fig. 5-38. This tube, too, is oftentimes called a triode, despite the fact that, as in the preceding case, four distinct elements are contained within the gun. The assembly differs from the previous one in the interchange of the position of the apertured masking baffle in the anode cylinder. Previously, the baffle faced the screen grid; in this gun it is positioned at the end facing the fluorescent screen of the tube. Also, the shape of the screen-grid electrode has been changed to what is almost an apertured disk.

The function of the baffle positioned at the emerging end of the anode barrel is obvious; it is intended to mask a portion of the beam, and in that way to make the beam smaller in diameter. Several effects are achieved. One of these is that fringe electrons in the bundle, which were not properly focused by the focusing coil, will be masked off, thus tending to retain the sharpness of the spot. The second effect of interest to us is that the dimensions of the beam entering the deflection area are reduced; consequently, change in



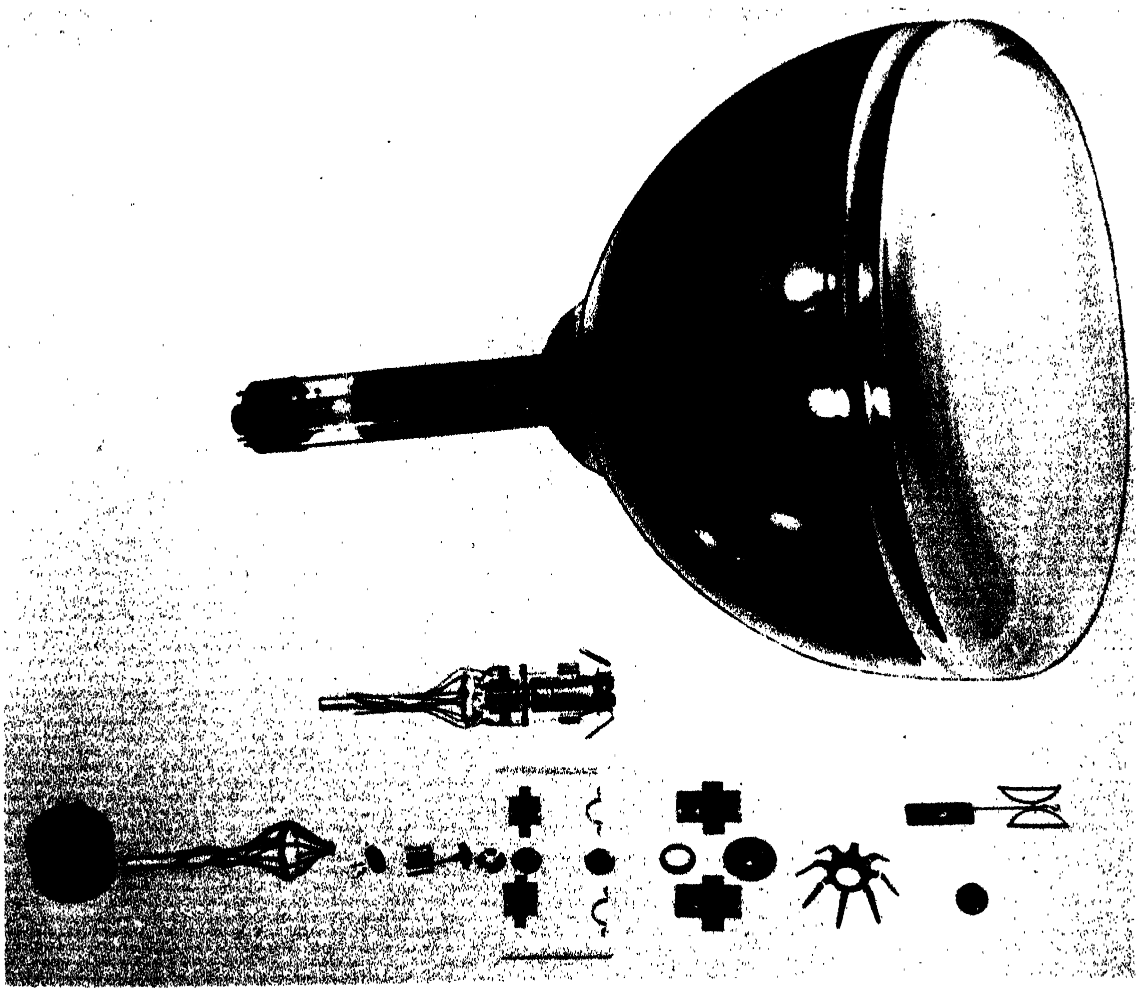
Courtesy Du Mont Labs.

Fig. 5-38.—Dimensional drawing of television-tetrode electromagnetic electron gun used in 12JP4, 15AP4, and 20BP4 cathode-ray tubes. Dimensions are in inches.

focusing, which might be introduced due to the deflection process, is minimized. A picture of a tube which employs this type of gun, along with the components of the gun and the assembled gun, is shown in Fig. 5-39.

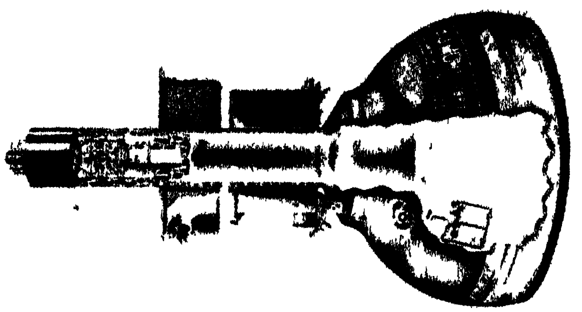
The organization of the component parts of the gun read from left to right, in the manner of an assembly diagram. The fact that each of the grids and the anode are formed from two half-sections is plainly visible, as are the apertured baffles which seem to be disks which are welded into position. Directly to the left of the control-grid half-sections, the cathode, the supporting framework, and the heater element are shown; the latter being closest to the stem assembly around which the wires are twisted. The long glass tube extending downward from the stem is used for evacuating the envelope and sealing the tube.

An idea of how such a gun appears inside of the envelope of a completed tube, including the relative posi-



Courtesy Du Mont Labs.

Fig. 5-39.—Finished tube, and components which make up electron gun of 12JP4 cathode-ray tube.



Courtesy Du Mont Labs

Fig. 5-40.—Cutaway view of 7BP7 tetrode electromagnetic cathode-ray tube.

tions of the focusing coil and the deflection coil, can be had from the cut-away view in Fig. 5-40. The peculiarly shaped structure which appears in the conical portion of the envelope, close to the accelerating anode connecting terminal, is the getter assembly. The getter is a material which is "fired" or vaporized after evacuation of the tube has been completed, and it serves to absorb the gases which otherwise might remain in the elements of the tube and later be freed during operation, causing improper performance.

Ion Bombardment

The liberation of electrons by the hot cathode is accompanied by the liberation of negative ions, which may also come from the elements around the cathode. Since even the lightest of ions is very much heavier than an electron, their presence poses a problem.

In electrostatic deflection of charged particles, the deflection angle is independent of the mass of the particle. Since electrostatic focusing is a deflection process, both ions and electrons will be brought to the same focus, and will also be deflected equally. Consequently, the beam tracing a pattern on the face of the tube will be composed of electrons and a small number of negative ions; these ions cause no undesirable results.

However, in electromagnetic fields, the deflection angle of a charged particle is inversely proportional to the mass of the particle. Because of the greater mass of an ion, those ions present in the electron stream will be neither focused at the same point as the electrons nor deflected by an appreciable amount. This will permit them to travel to and impinge on the center of the tube face, causing fluorescence, unless suitable means are taken to prevent such ion bombardment. If this bombardment is permitted to occur, "brown spots" or "ion spots" appear. Eventually the area covered by these spots becomes enlarged until finally the screen is useless for viewing; actually, even before the spot becomes excessively large, it is very much of a nuisance.

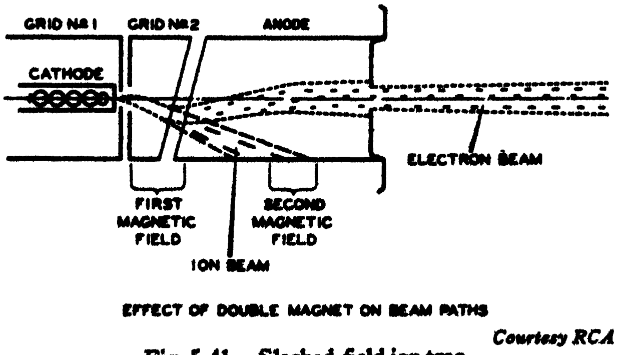
The removal of the ions from the beam can be accomplished by two methods involving the gun, and by a third method which prevents the detrimental effect of ion bombardment by a special treatment of the screen. This is a thin aluminum coating which is placed on the side of the screen facing the deflection area, and is explained later in connection with screens; only the first two methods will be discussed here.

Slashed-Field Gun with Ion Trap

The slashed-field gun is a modification of the physical construction of some parts of the tetrode gun. Although it retains the electrical organization and is still within the same basic category as the previous type, this gun introduces a decided change in the electron path.

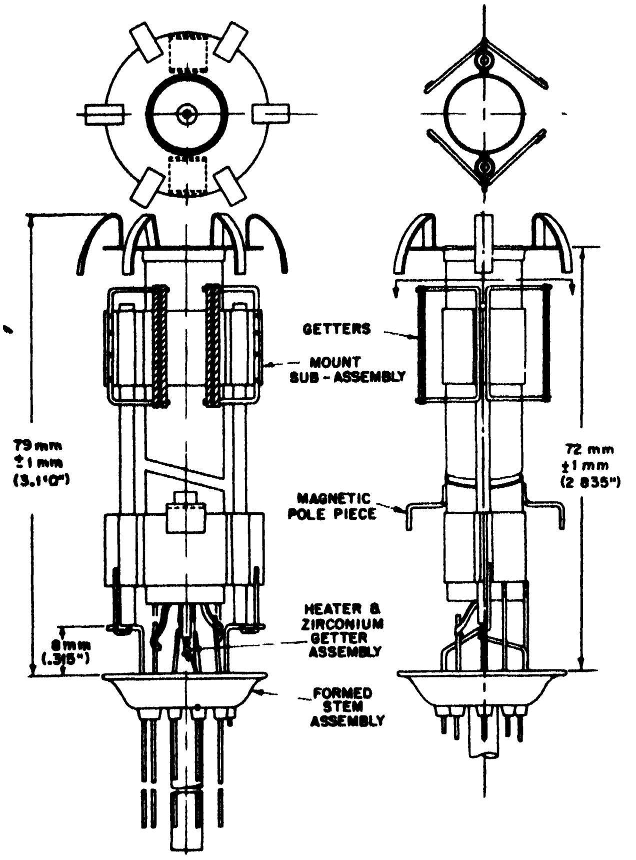
As shown in the detail drawing in Fig. 5-41, the cathode and control-grid electrodes require no comment. The screen-grid and anode electrodes, however, are peculiar in shape, the facing edges being cut diagonally at an angle of about 15°. Not shown in this drawing, but visible in Fig. 5-42, are two magnetic pole pieces or "flags" in the region of the slant lens. These serve to localize the field of a magnet which is placed around the neck of the tube at the time of installation.

The ion trap is created in the area occupied by the slant lens. By virtue of the slant in the facing edges of the screen grid and the accelerating anode, the electrostatic field formed there is tilted, thus creating a new axis for the system. Since the path followed by a charge is the same for an ion or an electron, both tend to follow a tilted path, according to the new axis. However, the action of a magnetic field on a charge is a function of the mass as well as the charge of the particle, so that by arranging a magnetic field in the region of the slant lens, the electrons are restored to the normal axial path. The ions, being virtually unaffected by the magnetic field of the ion-trap magnet, will be trapped within the accelerating anode barrel.



Courtesy RCA

Fig. 5-41.—Slashed-field ion trap.



Courtesy North American Philips

Fig. 5-42.—Complete slashed-field electron gun used in 10BP4 cathode-ray tube.

The two pole pieces, shown in Fig. 5-42, act in conjunction with the external magnet, which may be a permanent magnet, or may be operated from a d-c supply. Usually, a second magnet is used in this arrangement, being located about an inch closer to the screen than the first. Its function might be said to be twofold; first, to limit the field of the first magnet, and second, to aid in the proper centering of the image on the screen. The instructions for the adjustment of ion traps usually accompany the equipment. In most cases, such dual-magnet systems come as a single assembly. After being slipped onto the neck of the tube and brought to rest at certain markings on the tube, the ion-trap assembly is moved backward and forward slightly, as well as rotated a slight amount in each direction until the best trace is obtained on the screen. These comments are general and it certainly is recommended that the individual instructions for each equipment be followed in preference to these brief remarks.

Three types of ion-trap magnet systems have been used in television receivers, and probably in other applications as well. These are (1) dual magnets which receive their excitation from a d-c source, (2) dual permanent magnets, and (3) single permanent magnets. These are illustrated in Fig. 5-43.

Bent-Gun Ion Trap

Another version of the magnetic tube gun, intended to prevent ion bombardment of the screen, is shown in Fig. 5-44. The electrical organization and the number of elements are virtually the same as in the standard television tetrode. However, the accelerating anode is different in that it is bent, rather than straight. Part of the anode cylinder is coaxially symmetrical with the other elements, while the remainder is tilted away at an angle of 10.5°. It is this tilted axis which is on the main axis of the tube; the gun assembly is placed in the tube so that actually the cathode and grid are at an angle with respect to the tube axis.

Both ions and electrons enter the anode cylinder. But if a magnetic field is placed in the region of the cylinder, and so oriented that its field will exert a force on a negative charge in the direction of the bend, the electrons in the beam will be deflected along the main

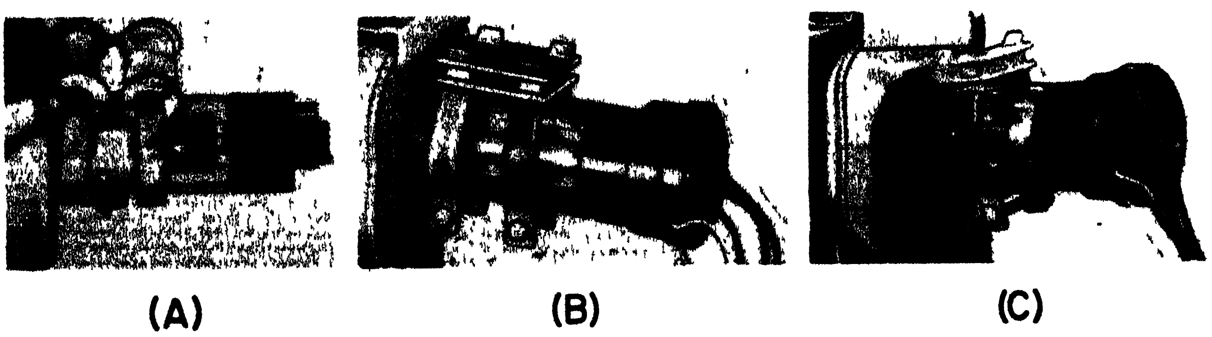


Fig. 5-43.—Typical ion-trap magnets: (A) dual electromagnet, (B) dual permanent magnet, (C) single permanent magnet.

two values, namely 6.3 volts at 0.6 ampere, and 2.5 volts at 2.1 amperes. Both of these values are used with both types of tubes, although the higher voltage rating is more common. Whichever is the actual voltage applied, the rated tolerance is plus or minus ten per cent, although it is highly recommended that it be as close to the nominal value as possible.

Grid Bias (Cutoff)

The exact relationships between the cutoff bias, screen-grid voltage, and the different anode voltages in electrostatic- and electromagnetic-type tube guns vary from type to type. It is possible to make general statements only if electrostatic tubes and electromagnetic tubes are considered separately. Although an analysis of this type may not be completely rigorous, it is sufficiently accurate to be acceptable.

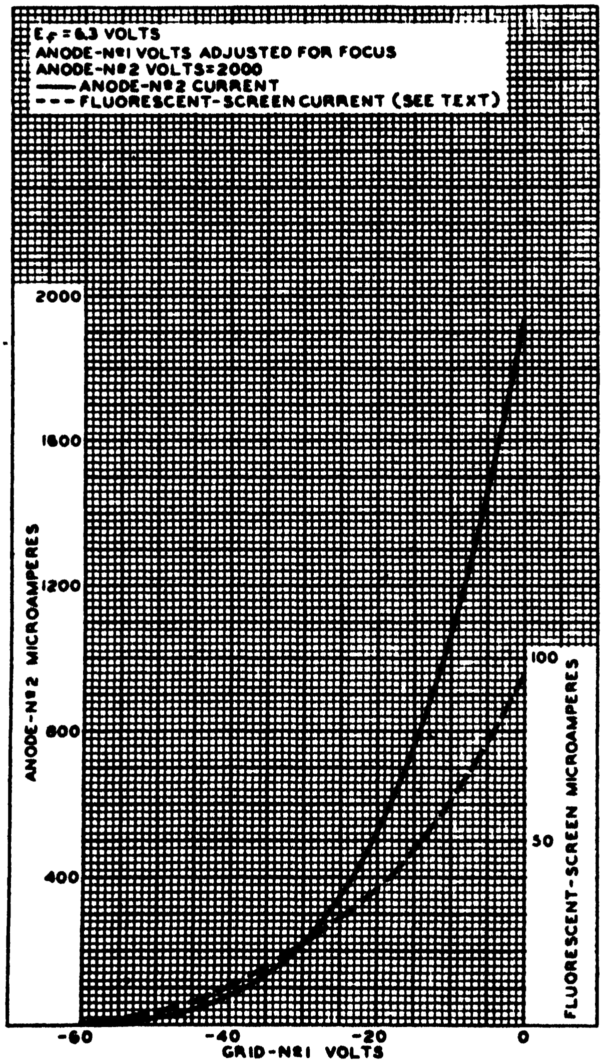
The cutoff bias voltage varies directly as the accelerating voltage in most electrostatic tubes, provided that the tube does not employ a separately connected screen grid. Such screen grids are not found in the electrostatic tubes shown herein. In electromagnetic tube guns, the cutoff bias varies directly as the screen-grid voltage, this electrode being separately connected in most magnetic tubes.

Grid Drive

A condition of operation associated with the grid is that which produces a stipulated trace intensity or light output at the screen when the anode voltages are of certain specified values. Sometimes the grid voltage is referenced to beam or screen current, rather than light output. By grid condition in this case we mean the grid voltage above cutoff necessary to drive the tube to the stated beam current or to the stated light output. This is known as the *grid drive.*

Tube charts specify this quantity because it is of interest to the designer of equipment, especially when a signal voltage is to be applied to the grid so as to *intensity-modulate* the beam current. Intensity modulation is a change in the intensity of the beam current or light output from the screen, in accordance with the changes in signal voltage at the control grid. The limits of intensity modulation obviously are zero beam current as one extreme, and that which will produce the full trace intensity as the other.

The designer is concerned with the grid-drive requirements of a cathode-ray tube because it indicates the level of the signal necessary at the grid to overcome the fixed grid bias and produce the specified beam current. The signal source must furnish this signal output. The relationship between grid drive and beam current, or light output, is known as the *grid-drive characteristic.* This is usually shown in graphic form, an example of which appears in Fig. 5-49.



Courtesy RCA

Fig. 5-49.—Grid-drive characteristics of 5UP1 cathode-ray tube.

In this case, two different conditions are indicated. One shows the grid-drive characteristic relative to second-anode current; the other curve shows the grid-drive characteristic relative to the screen current. It is interesting to note that at zero grid voltage, the ratio between screen and second-anode currents is almost 1 to 20, and while this ratio is not entirely constant throughout the range of grid voltage, it is substantially so. At any rate, it is evident of a previous statement concerning electrostatic focus that the charges which advance to the screen are comparatively few with respect to those which form the anode currents.

The grid-drive figures quoted in tube tables are average values representing measurements on a number of tubes of like type. A tube-to-tube variation should be expected because of manufacturing tolerances. Since many tubes do not permit the measurement of screen current, grid-drive characteristics are frequently referenced to the light output and identified as *modulation characteristics*. In turn, light output is expressed in foot-lamberts and is determined by developing a trace on the screen in the form of a number of horizontal lines — a raster two inches by two inches — and measuring the light output by means of a suitable meter. Standard methods of measurement of cathode-ray-tube constants are given in the Appendix.

High-Voltage versus Low-Voltage Tubes

While operating voltages do not establish distinct tube types, there does exist a general classification of tubes based upon the voltage applied to the accelerating anode. Generally speaking, electrostatic tubes operated with second-anode potentials of 2,000 volts or less are considered to be *low-voltage* tubes, whereas tubes with second-anode voltages in excess of this figure are classed as *high-voltage* tubes. The voltages applied to any intensifier anodes present are not considered; it is strictly the second-anode voltage which determines the general classification.

No one relationship affords advantages only; every advantage of one kind usually is accompanied by a disadvantage of some other kind. The low-voltage tube inherently has high sensitivity because of its low-velocity beam. The low-velocity beam, however, produces a trace of relatively low luminosity, a trace which may not be readily visible in a brightly lighted room, especially if the ambient light falls upon the face of the screen.

At first thought this may seem to reflect adversely upon the utility of the low-voltage cathode-ray tube. Such criticism is unwarranted, because the degree of brightness attainable with any low-voltage tube is adequate for many uses. In fact, the statement made is more true in theory than in practice for a number of reasons. First, while it is true that the trace is not sufficiently bright to be studied critically under the light conditions permissible with high-voltage tubes, the average oscilloscope containing a low-voltage cathode-ray tube will produce a trace which is perfectly readable when viewed under semishaded conditions. Second, present-day film permits photographing of this trace with very little difficulty. As a matter of fact, the majority of the oscillograms contained in this text were made with low-voltage cathode-ray tubes. Third, the advantages of high sensitivity are much more important for everyday oscilloscope applications than high trace intensity.

The high-velocity beam tube, however, is not without its advantages. Some conditions of operation demand viewing in broad daylight, which calls for a brilliant trace. This is also true if projection of the image is desired for classroom demonstration or television projection systems. High writing speeds require a bright trace because the higher the velocity of the electron beam, the less the time spent by the beam at any given point on the screen. Therefore, proper display demands high velocity of the beam, which means high voltage on the accelerating anode.

The higher the velocity of the beam electrons, the less is the possibility of deflection by extraneous fields such as those due to near-by operating transformers, chokes, etc. In addition, higher operating voltages result in a narrower beam and smaller spot. These pros and cons of the low-voltage and high-voltage tubes are theoretical generalities, usually found in practice, but not necessarily rigid rules of application. Most cathode-ray tubes are sufficiently flexible in their use to allow their application to most needs, even though those needs may be better satisfied by some other tube with more suitable operating parameters.

The last statement does not suggest that tubes be operated at voltages differing from those specified by the manufacturer, or that tubes especially developed for certain applications are suitable for indiscriminate use. It simply means that many requirements, which may be better satisfied with high-voltage tubes, can nevertheless be satisfactorily accomplished with low-voltage tubes, and vice-versa. The governing factor is the permissible departure from required performance.

Ratio of Anode Voltages in Electrostatic Tubes

Some degree of consistency exists between the focusing-anode voltage E_{b1} and the accelerating-anode voltage E_{b2} because of similarities in electron-gun design. The voltages applied to the individual electrodes in the different types of tubes are not necessarily the same, but as a rule, the ratio lies between 1 to 2 and 1 to 3. However, in some instances the second-anode voltage may be as much as four times as high as the focusing-electrode voltage. When both voltages are unknown, familiarity with this ratio may not be of much benefit, but if either of the voltages is known, these ratios may serve as a guide in the values of voltage which may be tried on the other electrode.

A range of ratios may be found between the voltages applied to the accelerating anode and the intensifier or post-deflection accelerating electrodes. Usually the ratio is about 2 to 1 with the higher voltage on the intensifier anode. In some tubes, where *very* bright traces are needed, the ratio may be as high as 4 to 1. If the intensifier voltage is not known, and since the post-deflection accelerating anode voltage always is higher than the second-anode voltage, a minimum value of twice this figure is safe to apply to the intensifier anode. If still greater trace brilliance is needed, higher voltages may be tried, although it must be remembered that the safe intensifier-anode voltage value is not unlimited.

This limitation exists for two reasons. First, the gap between the second-anode coating and the intensifier-anode coating may break down if the voltage applied to the latter anode exceeds the specified maximum. Second, with regard to the performance of the tube, if the intensifier voltage is too high the spot may be distorted. Another possible condition is "folding back" of the spot onto the screen after it has been deflected off the screen. It must be emphasized that the high accelerating voltage requires that the beam not be permitted to remain stationary on the screen, unless the intensity is reduced almost to total extinction of the beam. If the high-velocity beam is allowed to remain stationary on the screen, it will burn a spot on the screen coating and irreparably damage the screen.

There is an exception to the ordinary limits of the ratio between the second-anode and intensifier-anode voltages. This is the intensifier-anode voltage applied to the DuMont 5RP1, where maximum voltage rating of the intensifier anode is about ten times that applied to the second anode.

As far as general practice is concerned, the voltage applied to the intensifier anode should be as low as possible, consistent with the attainment of the desired trace brilliancy.

Spot-Size Considerations

The factors which control spot size are numerous, and will, of course, influence the width of the line traced on the screen. In regard to this relationship, it is noteworthy that the width of the line on the screen, assuming constant potentials on the various electrodes, will generally be thinner than the diameter of the stationary spot. The reason for this is that the current density in the spot is not uniform, being less toward the periphery. However, when the spot is stationary, the period of screen excitation for all parts of the spot is greater than when the spot is moving. Therefore, the eye sees substantially constant excitation of the screen for the full diameter of the spot. When the same spot is in motion, the greatest screen excitation will develop from that part of the spot which has the greatest current density, the center. Thus the excitation from the extremes of the spot is reduced and the line width is decreased. This condition, in turn, is affected by the speed of the spot as it moves across the screen; the faster the motion the less the width of the spot, and the less its brilliancy, since the excitation time is reduced.

Another item of interest, in connection with the relationship between the spot size and the line width, is the change in line width with the position of the spot on the screen. Assuming proper focusing adjustment and constant beam current, there is a tendency for the spot to widen as it advances from the center of the screen, where best focusing prevails, toward the edge. This is the result of two conditions, one of which is the curvature of the screen. The screen has a radius of curvature which is greater than the radius of curvature of the beam focal point locus. With the focusing adjusted for the center of the screen, the outer limits of the average screen are further from the focusing system than the center. Therefore, the focal point of the focusing lens is located at some point short of the screen, rather than at the distance corresponding to the screen location. Thus the beam strikes the limits of the screen, not at the point of minimum beam diameter, but after its second cross-over.

The second controlling agency is the deflection system; this subject will be described in Chapter 6.

General Conditions of Spot-Size Control

Increasing the beam current increases the spot diameter; increasing the accelerating voltage decreases the spot diameter, increasing the screen-grid voltage likewise decreases the spot diameter. These are generalities and are not intended to be applied to tubes of different design, because in many cases they will be found to conflict. Generally, exclusive of the projection-type tubes, the greater the screen dimension, the higher is the accelerating voltage, yet the spot dimension does not necessarily become smaller. Thus, the generalization is applied tube by tube, rather than tube against tube.

Spot Size versus Screen Diameter

At first thought, and logically so, one would imagine that the smallest spot is developed by the gun used in the tube which has the smallest screen. This is generally true, but not always. That it should be true is understandable, since the resolution of a picture depends upon the ability of the eye to separate two adjacent lines. Thus if the screen is smaller, the width of the line traced by the spot in that tube should be smaller.

This is true in the projection-type television tubes more than in any other. In these tubes, two of which are the 3NP4 and 5TP4, the spot dimensions are very small. In the 3NP4, the spot size is about 0.0023 inch, with a picture size of 1.42 x 1.89 inches. In the latter, the spot size is about 0.0035 inch, and the picture size approximately 2.55 x 3.4 inches.

As to other tubes, the following sample spot dimensions are of interest. Some electrostatic tubes with 3-inch screens have spot dimensions of 0.024 inch, 0.03 inch, and 0.04 inch, while one of them has a spot diameter of only 0.018 inch. Similar tubes with a 12-inch screen utilize a spot which is about 0.072 inch in diameter. Even these few figures are sufficient to show that the spot dimension is a function of the gun design, which in turn is related to the specific purpose of the tube. Discounting the special function tubes, it is safe to assume that the larger the screen diameter, the larger will be the spot diameter, although not necessarily in proportion to the increase in screen diameter.

That which is true about the electrostatic tubes will be found to be true in the magnetic type of tube, despite the general impression that since magnetic tubes operate at higher accelerating voltages, the spot will be very much smaller. The spot diameter in such tubes frequently is smaller, but again it is a matter of gun design and intended use of the tube. For example, and just as a matter of simple comparison, here are some spot diameters in magnetic-type tubes. Two typical 7-inch tubes function with spot diameters of 0.024 inch and about 0.034 inch; some 9- and 12-inch tubes have a spot which is about 0.06 inch in diameter. And a 4-inch tube has a spot that is 0.012 inch in diameter, whereas a 3-inch tube has a 0.024-inch spot.

Spot-Size versus Focus-Coil Position

It has been found that the position of the focus coil tends to affect the spot dimension; as the focus coil is moved toward the screen, the spot diameter becomes larger. After the current through the coil has been readjusted for proper focusing, moving the focus coil toward the gun causes distortion of the spot.[8]

It is also interesting to note that focus-coil gap has an effect on the spot size, tending to increase the spot dimension as the gap width is increased. This, however, is not very important to the general user, since focus coils used in television and many other systems are fixed in dimension. It is only when the coil is equipped with a variable gap that this consideration need be recognized.

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[8] Soller, T., Starr, M. A., and Valley, G. E., Jr., "Cathode Ray Tube Displays," 1st ed., pp. 96-98, McGraw-Hill Book Co., Inc., New York, N. Y., 1948.

CHAPTER 6

DEFLECTION SYSTEMS IN CATHODE-RAY TUBES

Two methods of deflecting electron beams are used in cathode-ray tubes. They are electrostatic and electromagnetic fields. The principles underlying these actions were explained in Chapters 2 and 3, hence need not be repeated here.

ELECTROSTATIC DEFLECTION SYSTEMS

Electron Velocities in Cathode-Ray Tubes

Inasmuch as the discussion to follow will frequently involve references to electron velocities in one way or another, it might be well to devote a few paragraphs to the subject. Already, we have shown how the velocity of an electron is a function of the accelerating potential, varying as the square root of the potential, but it will be better if we resolve velocity in terms of actual time intervals.

Electrons have been accelerated to almost the speed of light, approximately 299,800,000 meters per second, but cannot equal it, because, according to Einstein, the mass of the electron becomes infinite at the speed of light. Examining the general run of cathode-ray equipment, we find that accelerating voltages employed in such apparatus range from about 1,000 volts as the low limit to about 50,000 volts as the high limit. Over this range of acceleration, the electron velocity approximates about 6.25 per cent of that of light at the low end, to about 44 per cent of the speed of light at the high end. These velocity figures already include the correction factor which must be applied to the mass of the electron as the accelerating voltage increases about 10,000 volts.

If we use an accelerating voltage of 1,000 volts as the basis of this discussion, the velocity of the electron is calculated as being $0.0625 \times 3 \times 10^8$ or 18,750,000 meters per second, which is equal to 18.75 meters per microsecond or 739 inches per microsecond. This corresponds to a time interval of 1,352 micromicroseconds per inch (2.54 cm) of travel. At an accelerating potential of 50,000 volts, the same distance would be covered seven times more rapidly or in 191 micromicroseconds.

If we reduce our dimensions still further, let us say 0.1 inch, the 1,000-volt beam will traverse this distance in 135 micromicroseconds, and the 50,000-volt beam will span it in 19.1 micromicroseconds; hence the two beams will cover a distance of ¼ inch in 338 and 48 micromicroseconds, respectively.

Rectangular and Polar Co-ordinates

Electrostatic deflection systems may be of different types, although the fundamental principles underlying their operation remain the same. The final display of information on the cathode-ray-tube screen may be organized differently in different kinds of oscilloscopes. By this we mean that the final image which appears on the screen may be shown in what is known as *rectangular co-ordinates,* or in the so-called *polar co-ordinates.* In the rectangular co-ordinate system, the quantities involved are shown along two axes or co-ordinates which are mutually perpendicular. In the *polar coordinate* system, the quantities involved are shown along a *path* described by a rotating radius of varying length.

The customary manner of illustrating a periodic quantity such as a sine wave, complex wave, or a resonance curve, is representative of the use of rectangular co-ordinates, which are mutually perpendicular, as indicated in Figs. 6-1A through C. The difference between these three drawings is simply the point where the two co-ordinates intersect each other. The co-ordinate or axis which is oriented perpendicular to a reference horizontal axis such as the earth's surface always is identified as the *Y* axis or *Y* co-ordinate; the co-ordinate or axis which is oriented parallel to the reference horizontal axis is known as the *X* axis or *X* co-ordinate. *Usually* amplitude is shown along the *Y* axis, and time is shown along the *X* axis, but this type of organization is not a must; any combination of functionally related quantities may be shown, as indicated in Figs. 6-1D through F. Fig. 6-1D illustrates the response curve of a discriminator system; Fig. 6-1E shows a conventional sine wave and Fig. 6-1F illustrates a tube characteristic. All of these are traces which can be displayed on a cathode-ray-tube screen, and are but a few of the very many different types of rectangular co-ordinate presentations.

In every instance one quantity is shown along the vertical or *Y* co-ordinate and the other quantity, whatever it may be, is shown along the *X* axis. In Fig. 6-1D, output voltage is shown on the *Y* axis and frequency is

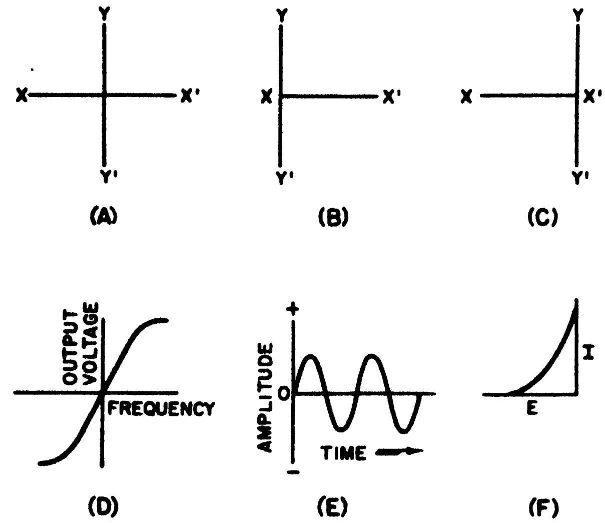
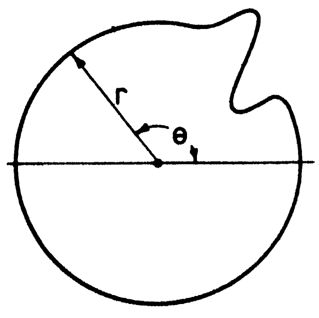


Fig. 6-1.—The conventional rectangular co-ordinate system shown in (A), while (B) shows rectangular representation utilizing the *Y*-axes and the positive *X*-axis, and (C) represents the *Y*-axis and the negative *X*-axis. (D), (E), and (F) illustrate typical rectangular representations of functionally related quantities, such as the response curve of a discriminator system, an ordinary sine wave, and a triode tube characteristic, respectively.

shown on the *X* axis; in Fig. 6-1E, amplitude appears along the *Y* dimension and time is shown along the *X* dimension, the latter being graduated in time, electrical degrees, radians, or whatever other units are used to express time. In Fig. 6-1F, the *Y* axis indicates current and the *X* axis is calibrated in terms of voltage.

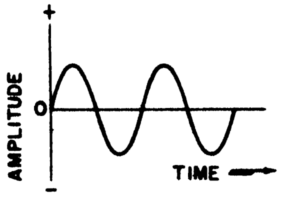
The polar co-ordinate method of presentation can most easily be represented by a circular trace as in Fig, 6-2, wherein the circular path is traced by the radius *r* rotating at a constant angular velocity. In the illustration shown, a sine wave is presented along polar co-ordinates. More detailed examples of this type of display will be given later.

Fig. 6-2.—The polar method of representation illustrates the relation between a variable radius *r* rotating at constant angular velocity and the angle *θ* which this radius forms with a reference line, here taken along the horizontal. A sine wave, displayed by this method, is shown in the figure.



In the meantime, reference to the illustrations in Fig. 6-1 and in Fig. 6-2 makes it clear that the presentation of information concerning a quantity which varies relative to some other factor requires a two-dimensional display. In other words, anything which involves two variables can be displayed properly only in two dimensions. For example, the sine-wave curve in Fig. 6-1E, repeated as a matter of convenience in Fig. 6-3, would be a straight line if the time dimension were omitted, and as such would furnish very limited information,—in this case only the over-all peak-to-peak amplitude, if suitably calibrated. Any information concerning the instantaneous changes with time would be absent.

Fig. 6-3.—The sine wave shown in Fig. 6-1E is displayed along rectangular co-ordinates.



The Two-Dimensional Deflection System

In order to produce a display along rectangular co-ordinates, the standard electrostatic tube employs two deflection systems. Each of these consists of a pair of plates so arranged relative to each other that the fields created by voltages applied to each pair are positioned at right angles to each other. In this way, one field produces a deflection along the *Y* axis and the other field produces a deflection along the *X* axis. Each can be used independently of the other or they can be used simultaneously.

Since each of the deflection fields is assumed to be uniform, the lines of force of the two fields are mutually perpendicular. A pictorial representation of these two fields looking head-on from the front of the tube would look like Fig. 6-4. Everywhere the lines of force of one field are at right angles to the lines of force of the other field. The pairs of plates responsible for these fields are *A-B* and *C-D*.

Understandably, the orientation of the two deflection systems is relative to the reference plane. Since the fundamental principle of operation is independent of the physical orientation of the plates, either set of plates may be used to produce a deflection in either direction. It is simply a matter of how the tube as a whole and hence the plates within the envelope are positioned.

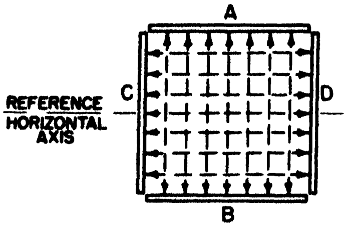


Fig. 6-4.—Imaginary top view of the crossed field lines produced by the vertical-deflection plates *A-B* and the horizontal plates *C-D*.

Turning the tube through 90° in either direction inverts the relative orientation of the two deflection systems.

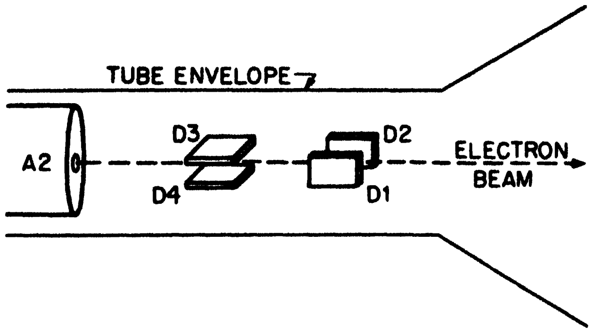


Fig. 6-5—The arrangement of the deflection plates within the tube; one pair of plates is closer to the accelerating anode *A2* than the other, assuring independent beam deflection by each set of plates.

Constants of Electrostatic Deflection Systems

In order to analyze the constants of electrostatic deflection systems, it is best to examine each system separately. So, for the moment, we shall assume that only one set of plates exists within the envelope; the pair which is closest to anode 2. To most conveniently attain our goal, we shall once more discuss the fundamental action taking place within a pair of charged surfaces, except that this time we shall associate the action with a number of other variables. It should be understood that such a one-dimensional deflection system is not entirely useless, but from the modern viewpoint of what a cathode-ray tube should do, it is impractical. Nevertheless, it does serve as a basis of analysis.

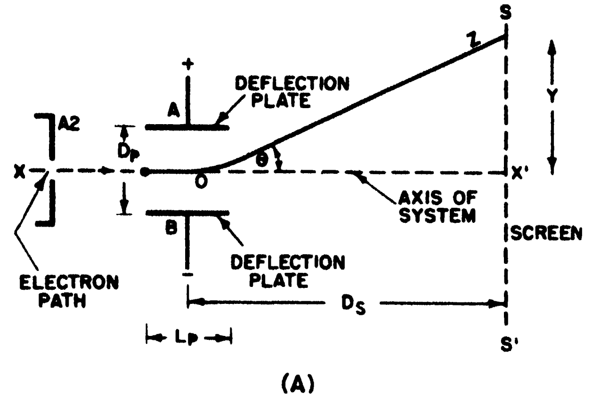
The composite field pattern shown in Fig. 6-4 may seem to contradict the earlier statement that electrical lines of force cannot cross each other. The statement remains valid, and the contrary impression stems from the manner of illustrating the positions of the two deflection systems. Actually they are not located in the box-like manner indicated in the figure; instead they are side by side along the path of the electron. In other words, the two deflection systems are laterally displaced relative to the origin of the beam, that is, the aperture in the accelerating anode. The box-like arrangement shown in Fig. 6-4 was tried by Von Ardenne, but was found wanting, because the single field produced by the two voltages was not uniform throughout the space between the two sets of plates.

In present-day electrostatic tubes, the two deflection systems are arranged as shown in Fig. 6-5. Exceptions to this arrangement exist, and will be described, but for the present, we can forget about them because the method shown is standard in better than 99 per cent of the electrostatic tubes in use. One set of plates is closer to the accelerating anode *A2*, than is the other. Thus each field acts *separately* on the beam, and *theoretically independently* of the other. Since the beam may be viewed as originating at *A2*, the deflection system closest to that electrode acts on the beam first, followed a short time later by the second deflection system. For the moment, we can assume that this lapse of time is so short as to be entirely negligible, so that in effect, both plates act in concert. Be that as it may, the usual circuit arrangement is to employ the deflection system nearest the electron gun (anode 2) to produce the vertical or *Y* deflection, and the deflection system farthest removed from the gun, to produce the horizontal or *X* deflection.

In Fig. 6-6A the dotted line *X-X′* is the axis of the tube system. The electrode *A2* is the apertured accelerating anode operated at a potential V_a. The electron beam produced within the gun issues through the aperture and enters the deflection area with a velocity determined by V_a. The two horizontal plates (vertical-deflection plates) are a distance D_P apart and each plate has a length L_P. The two plates are subjected to a difference of potential (the deflection voltage) V_p. The electrostatic field present between plates *A* and *B* is a function of V_p, and its intensity may be expressed by

$$E=\frac{V_p}{D_p}. \qquad (6\text{-}1)$$

It is significant to note that the midpoint of the deflection system is at a distance D_s from the screen *S-S′*. This, then, is the geometry of the single-element electrostatic deflection system. Let us now assume that the deflection voltage is temporarily removed, and the electron beam, propelled into the deflection area, advances along the axis *X-X′* unimpeded until it arrives at the screen, and strikes it at *X′*, the geometric center of the screen surface. Now the deflection voltage is applied, and according to the principles described in Chapter 2, an electrostatic field is created within the space between plates *A* and *B*. Because of the polarity of the



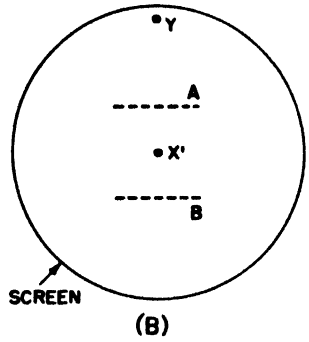


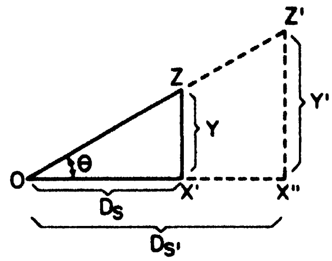
Fig. 6-6.—(A) is a side view of the deflection system showing the important geometrical constants affecting the deflection sensitivity, and (B) is a head-on view of the screen, showing the deflected spot position *Y*.

applied voltage, the lines of force are directed toward plate *A*. This will deflect the beam upward causing it to leave the deflection area (theoretically limited by the physical boundaries of the two surfaces) at an angle θ relative to the axis *X-X'*. The beam now strikes the screen at a point *Z*, which corresponds to a displacement *Y* from the geometric center *X'*. This change in the location of the fluorescent spot on the face of the tube would be as shown in Fig. 6-6B.

Continuing with the analysis of Fig. 6-6A, let us examine some of the variables. What determines the distance *Y* — or the deflection *on the screen?* The first impression is that it is a function of the angle θ. That is correct, but in turn, this angle is determined by four other related conditions. These are the accelerating potential V_a, the deflection voltage V_p, the length L_p of the two plates, the separation D_p between them, and the distance D_s between the deflection plates and the screen.

The reasons for the control of the deflection angle θ by these factors is simple. The higher the accelerating voltage V_a, the more rapidly will the beam traverse the field space. Therefore, the less time will be available for the deflection force to act on the beam. Consequently, angle θ will be reduced. This is so regardless of the value of the deflection voltage, the length of the plates, or the separation between them, all of which are assumed constant in value during the variation of V_a.

Fig. 6-7.—The dependence of the vertical deflection on the distance D_s between the screen and the deflection center *O* for constant deflection angle θ.



For any one value of V_a, that is, beam velocity, the higher the deflection voltage V_p, the greater will be the angle θ. The same is true, if V_a and V_p and the plate separation D_p are maintained constant, and the length L_p is increased; or if V_a, V_p, and L_p are constant and D_p is *decreased*.

Now the path taken by the beam after deflection relative to the undeflected path *X-X'*, forms the angle θ, and if we project these two imaginary lines — the undeflected beam path and the deflected beam path to the screen surface,— we have formed the right triangle shown in Fig. 6-7. The base of this triangle is the tube axis *O-X'*, the hypotenuse *O-Z* (the deflected beam path) and the altitude is *Z-X'*, which equals the displacement of the spot along the screen surface. By reference to Fig. 6-6A, we see that the base of this right triangle corresponds to D_s, the distance to the screen from the midpoint of the deflection plates.

Without recourse to any mathematics, it is readily evident that if the distance D_s is increased to become D_s' in Fig. 6-7, shown by the dotted-line extension, the deflection angle θ remains as before, but the beam path will be increased to *O-Z'* and it will strike the screen at *Z'*, a greater distance from *X'* than before, thus increasing the deflection to *Y'*.

Mathematically, the dimension of *Y* is expressed by

$$Y = D_s \tan \theta \tag{6-2}$$

and

$$Y' = D_s' \tan \theta. \tag{6-3}$$

From the foregoing, it is evident that the deflection of the spot on the screen for any given value of deflection angle is determined by the *distance between the center of the deflection plates and the screen.* The greater this distance for any fixed operating condition, the greater is the deflection on the screen. This is true regardless of the direction in which the field is active, and regardless of the orientation of the deflection plates.

From this, one might arrive at the general conclusion that the longer the over-all length of the cathode-ray tube, the greater will be the deflection per unit voltage applied to the deflection plates. Generally speaking, this is true, if the tubes being compared are of like types, as far as accelerating potentials are concerned. For any comparison of tubes, all of the variables must be taken into account. Differentiation between tubes relative to meeting certain specific requirements is made easy by certain constants. Among these are deflection sensitivity and deflection factor ratings.

Deflection Sensitivity

Taking electrostatic tubes as a general category, the deflection on the screen of any one tube is determinable from the equation

$$\text{deflection sensitivity} = \frac{D_s \times L_p \times V_p}{2D_p \times V_a} \text{ cm per volt d.c.} \qquad (6\text{-}4)$$

where all lengths are in centimeters (1 cm = 0.3937 inch) and the potentials are in practical units of volts.

From this we can see that if all but one factor at a time are held constant:

(1) the deflection is proportional to
  (a) the distance between the deflection plates and the screen D_s
  (b) the length of the deflection plates L_p
  (c) the deflection voltage V_p;
(2) and the deflection is inversely proportional to
  (a) the separation between the deflection plates D_p
  (b) the accelerating voltage V_a.

Since the screen usually is described in linear dimensions, as for example, a 3-inch, a 5-inch screen, etc., by which is meant the full diameter (not necessarily the useful diameter), one might expect that the dimensions of the image formed on the screen would also be described in inches. Actually that is not the case; it is customary to state deflection in units of the metric system, usually in millimeters, where 10 mm = 1 cm = 0.3937 inch. However, the mere reference to a certain number of millimeters of Y-deflection would be meaningless, unless the basis of the deflection also were given because, as we have seen, many factors control the extent of the deflection. Foremost among these is the deflection voltage which determines the utility of the tube for certain applications. An image of suitable size can be obtained only if the relationship between the deflection voltage and the consequent deflection on the screen permits it. There are also other factors which govern the suitability of the tube, as for example, the intensity of the trace, but for the moment, we will concern ourselves with the dimensional aspects. The relation between the deflection voltage and the deflection on the screen is the basis for the ***deflection sensitivity.***

This constant determines the size of the image, i.e., the deflection of the spot per volt of deflection voltage. For example, tube specifications may state a "deflection sensitivity" of 0.2 mm per volt d.c. By this is meant that, when operating the tube according to the stipulated conditions, every 1 volt of d.c. applied to the deflection plates will cause a deflection of 0.2 mm on the screen relative to the undeflected position of the spot. Some other tube may state that its deflection sensitivity is 0.5 mm per volt d.c., which shows that the second tube will produce a deflection two and one-half times larger for the same value of voltage on the deflection plates.

As an example of the above, let us assume a hypothetical tube in which

$D_s = 20.32$ cm (8 inches)
$D_p = 1$ cm
$L_p = 2.54$ cm (1 inch)
$V_p = 1$ volt d.c.
$V_a = 1{,}000$ volts d.c.

We determine the deflection sensitivity by substituting the above values in equation (6-4). The equation now reads

$$\text{deflection sensitivity} = \frac{20.32 \times 2.54 \times 1}{2 \times 1 \times 1{,}000}$$
$$= 0.0258 \text{ cm per volt} = 0.258 \text{ mm per volt.}$$

In the event that the sensitivity is expressed in some other linear dimension, as for example inch per volt d.c., the significance of the term is not altered for comparing sensitivities. It is simply a matter of converting the deflection dimensions into like units.

Deflection Factor

The deflection factor is another commonly used term associated with the behavior of the tube, and indicates the voltage ***required*** on the deflection plate to produce unit deflection on the screen. This too is a basis of comparison and it is valuable for the individual who designs the equipment which will furnish the deflection voltage to the tube. Both ratings are to be found in tube specifications.

Expressed mathematically, deflection factor is the reciprocal of the deflection sensitivity, or

$$DF = \frac{2D_p \times V_a}{D_s \times L_p} \text{ volts d.c. per cm} \qquad (6\text{-}5)$$

and for the typical tube constants previously used, the converted ratio would be

$$DF = \frac{2 \times 1 \times 1{,}000}{20.33 \times 2.54}$$
$$= 38.77 \text{ volts d.c. per cm}$$
$$= 98.4 \text{ volts d.c. per inch.}$$

Some tube specifications define the deflection factor in terms of the second-anode voltage, especially when the tube is of such type as to be operable over a range of accelerating voltages. The meaning of the rating does not change from that which we have explained. For example, a tube usable with 1,500 volts or 2,000 volts on anode 2 may be rated on the basis of 1,000 volts, the recommended minimum voltage applicable to anode 2, in which case, the deflection factor is stated as follows

$D1$-$D2$ 68-92 volts d.c. per inch/kv of E_{b2}
$D3$-$D4$ 50-68 volts d.c. per inch/kv of E_{b2}

where $D1$-$D2$ are the pair of deflection plates closest to the screen, and $D3$-$D4$ are the deflection plates located closest to the electron gun. All of this means that this range of deflection factors is multiplied by the ratio $E_{b2}/1{,}000$.

Thus if E_{b2} is 1,500 volts, the deflection-factor range increases by 50 per cent to become

$D1$-$D2$ 102-138 volts d.c. per inch
$D3$-$D4$ 75-102 volts d.c. per inch

and with 2,000 volts on anode 2, the ratings increase to

$D1$-$D2$ 136-184 volts d.c. per inch
$D3$-$D4$ 100-136 volts d.c. per inch.

Such specifications also indicate the range of voltages applicable to anode 1, since the voltage applied to this electrode also has a bearing on the over-all operation of the tube. The lower limit of the first-anode voltage corresponds to the lower limit of the deflection factor at any specified second-anode voltage. Examples of this appear later in this chapter.

Deflection Factor versus Deflection Sensitivity

Two ways of expressing cathode-ray-tube sensitivity have been described. It is seen that they vary in opposite directions; when the deflection sensitivity goes up, as shown by the highest mm per volt rating, the deflection-factor rating goes down. The most sensitive tube has the lowest volts per inch rating. When two tubes are compared on the basis of deflection factor, the tube with the lowest volts-per-inch rating is the more sensitive tube.

Concerning the deflection sensitivity and the deflection-factor ratings, the direction of the deflection does not in any way alter the rating of the tube. The example chosen showed upward deflection of the spot. This was arbitrary and could just as readily have been a downward deflection. It is to be remembered that the character of the voltage applied to the deflection plate does not influence the sensitivity; the geometry of the tube and its operating potentials do that. The voltage applied to the deflection plates is not an operating voltage.

A-C versus D-C Deflection Voltage

Another interesting point is the relationship between the d-c sensitivity rating and an a-c voltage applied to the deflection plate. Assuming this to be a low frequency, say 60 cps, the d-c deflection shown in Fig. 6-6A corresponds to the *peak* voltage of the *positive* half of the cycle. If a 60-cps voltage were applied to the plates shown in Fig. 6-6A, the trace on the screen would be a straight line having a length twice as long as the length Y, one half of this line being located above the point X', and the other half being located below this reference point. The application of an alternating voltage to the deflection plates causes each plate to assume a different polarity alternately, thus deflecting the electron beam upward and downward, with the maximum spot displacement being determined by the peak voltage values. This is shown in Fig. 6-8. Thus *half* of the complete line formed by an a-c deflection voltage corresponds to the displacement of the spot when a d-c voltage equal to the peak value of the a-c voltage is applied.

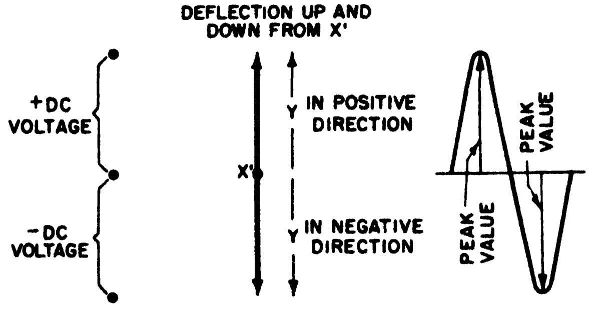


Fig. 6-8.—The deflection of the spot when an a-c voltage is applied to the vertical-deflection plates.

Two Sets of Deflection Plates

As we described earlier in this chapter electrostatic-deflection cathode-ray tubes utilize two sets of deflection plates which are located along the axis of the tube system, which is the path of the undeflected electron beam. Having determined the constants of one deflection system, let us now investigate the second system and its relationship to the first.

In Fig. 6-9 are shown two pairs of deflection plates. Disregard their shape because this simplified illustration is not an attempt to show the exact physical details of the deflection system. One pair of plates is

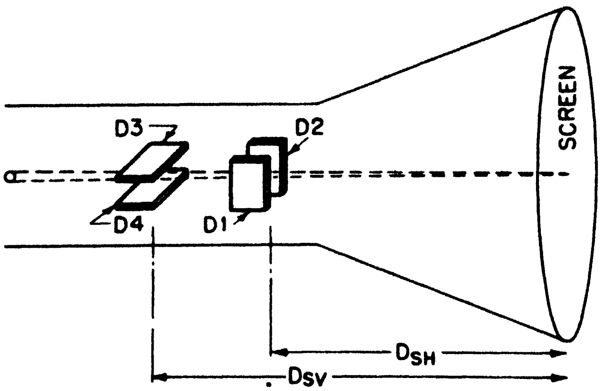


Fig. 6-9.—The arrangement of the deflection plates in the tube; note that the plates *D3-D4* are closer to the electron gun, and hence have higher deflection sensitivity.

identified as *D*3 and *D*4. These are located closest to the electron gun, and it is important to note that it is standard tube practice to identify the deflection plates *nearest the accelerating anode 2 by these designations.* In some instances, the letter *J* may be interposed between the *D* and the number. Throughout this book we omit the letter *J* and use but a single letter and its associated number. The *D*3-*D*4 plates are at a distance D_{sv} from the screen as shown in Fig. 6-9. The *D*3-*D*4 plates in a tube are usually the vertical deflection plates; however, in a few instances, it is recommended that they be used for the horizontal deflection.

The second set of plates, the plates nearest to the screen, is turned through 90° relative to the first, and usually produces the horizontal deflection. These plates are marked *D*1-*D*2 in accordance with standard tube engineering practice. If viewed from the front of the tube, the two sets of plates will present four edges as shown in Fig. 6-10.

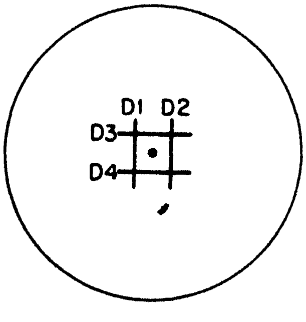


Fig. 6-10. — Head-on view showing the conventional arrangement of the *D1-D2* and the *D3-D4* plates; the *D3-D4* plates are usually the vertical-deflection plates, while the *D1-D2* plates perform horizontal deflection. A 90° rotation of the tube, however, will interchange their function.

The deflection sensitivity of the *D*1-*D*2 pair of plates which is closest to the screen will be less than for the *D*3-*D*4 plates, if for no other reason than that a projection of a unit deflection angle to the screen will produce a smaller deflection because that surface is closer to the deflection system. In turn, the deflection-factor rating will be higher. This condition will be found to be true in all electrostatic-type tubes containing a dual deflection system, such as we have described, and it is so shown in tube specifications.

Changing the orientation of the tube so that the directions of deflection are inverted does not alter the sensitivity associated with each set of plates. If the horizontal pair is used for vertical deflection by turning the tube through 90°, the sensitivity rating of that pair is still less than for the other pair of plates. In some instances, when such tubes are used as television picture tubes, the recommendation is made to turn the tube as a whole through 90° and to use the deflection plates nearest the gun for the horizontal deflection, and the other pair of plates for the vertical deflection. We recall this in connection with the 7DP4 tube. The reason for such a suggestion is that the ratio of sensitivities of the two deflection systems approximates the 4 to 3 aspect ratio of the picture. Thus equal values of signal voltages may be applied to both sets of plates and still produce the properly dimensioned raster on the screen. Normally, the horizontal deflection system demands a higher signal voltage since the picture dimension is longer in that direction.

Vertical versus Horizontal Deflection Sensitivity

The fact, that the horizontal deflection system in a cathode-ray tube is less sensitive than the vertical deflection system has been established. It is also of interest to know that the ratio between these two sensitivity ratings is not constant in all tubes of different type number and different over-all dimensions. It would be natural to expect a variation when the tube type numbers differed, but it would also be reasonable to imagine that tubes of like over-all length might have equal sensitivity ratings, or at least that the ratio between the ratings of these two deflection systems would be the same. That is not so in practice, however. In some instances, the difference in deflection-factor ratings for the two deflection systems may be as much as 50 per cent for some tubes of a certain over-all length, and in another tube of the same length, it may be only 12 to 15 per cent. Assumptions of any kind are unsafe, and it is recommended that tube specifications be consulted if comparative data are required.

Tube Length versus Electrostatic Deflection Factor

It might seem from Figs. 6-7 and 6-9 that the tube with the greatest over-all length would have the highest sensitivity or lowest deflection factor. Generally speaking, one might say that such reasoning is sound, yet in practice, this should not be taken for granted, since other factors are also involved. Below are given some comparative figures for a few popular types of tubes.

TUBE TYPES

	3BP1A		3KP1		5BP1	
E_{b1}	430	575	160–300	320–600	337	450
E_{b2}	1,500	2,000	1,000	2,000	1,500	2,000
DF (*D*1-*D*2) v/in.	150	200	50–68	100–136	63	84
DF (*D*3-*D*4) v/in	111	148	38–52	76–104	57	76
Over-all length in.	10¼	10¼	10¼	10¼	16¼	16¼

For equal second-anode voltages, the longer tube is usually the more sensitive one. However, the 3KP1, a tube only 10¼ inches long, operated with 1000 volts on anode 2 is more sensitive than any of the longer tubes listed in the table.

Effect of Intensifier Voltage on Deflection Factor

Earlier in this text (in Chapter 5) we discussed the pros and cons of post-deflection acceleration. Here are some figures to show that this form of trace intensification affects the sensitivity rating of the tube, but not to such an extreme as to defeat its usefulness. In fact the sacrifice in sensitivity is relatively low, and in numerous applications where high trace intensity is essential to successful application of the tube, the advantages far outweigh the disadvantages.

TUBE TYPES

	3JP7			5CP1		
E_{b3}	2,000	3,000	4,000	2,000	3,000	4,000
E_{b2}	2,000	1,500	2,000	2,000	1,500	2,000
E_{b1}	400–690	300–517	400–690	400–690	300–517	400–690
DF (*D*1-*D*2)	138–184	127–173	170–230	62–84	59–80	78–106
DF (*D*3-*D*4)	100–136	94–128	125–170	54–74	50–68	66–90

Examining these figures, we note that the deflection factor is increased approximately 25 per cent as a consequence of making $E_{b3} = 2\,E_{b2}$ relative to $E_{b3} = E_{b2}$.

Practical Aspects of Sensitivity Ratings

From the practical standpoint, deflection sensitivity and deflection factor mean certain specific things. To begin with, we have thus far only discussed the tube alone and ignored complete equipment, that is, amplifiers used ahead of the two deflection systems. Therefore, a difference in deflection-factor rating means that for *like* values of signal voltage applied to the two sets of plates, the deflections obtained will have different values. The deflection system with the highest sensitivity rating (deflection-factor rating) will produce the trace of largest dimension.

Naturally, if unlike voltages are applied to the two deflection systems, and their ratio is exactly the reciprocal of the ratio of the two deflection-system ratings, equally deflected traces will be produced by both systems. Thus if the horizontal deflection system is rated 50 per cent below the vertical deflection system, and the peak value of the voltage applied to the horizontal system exceeds the peak value of voltage applied to the vertical system by 50 per cent, equal traces will be produced by both deflection systems.

When amplifiers are used ahead of the two deflection systems, the tube ratings can retain their significance, or they can be nullified completely, since the contribution made by the amplifying system can more than offset any difference in the tube ratings.

The use of amplifiers ahead of the two deflection systems modifies the over-all ratings, which invariably are given in terms of deflection factor. Now the size of the trace is determined by the basic sensitivity of the tube *and* the amplifying capabilities of the amplifier. Due to the latter, the voltage required at the input of the amplifier for a unit dimension trace is reduced far below that required at the terminals of the deflection plates. Moreover, since the amplifier is an a-c device, the deflection-factor rating now is given in terms of rms values instead of d-c values.

For example, the deflection factor of the 3AP type tube is 76 volts d.c. per inch on the horizontal deflection plates *D*1-*D*2 and 73 volts d.c. per inch on the vertical deflection plates *D*3-*D*4 with 1,000 volts accelerating potential. The complete equipment using the same tube and operated at an accelerating potential of 1,100 volts affords a deflection factor of 0.7 volts rms per inch for both the *X* and *Y* deflection systems. The increased sensitivity is due to the amplifiers which are associated with each of the deflection systems. This is explained more fully in Chapter 10.

Sometimes the deflection-factor rating of the horizontal deflection system in complete oscilloscope equipment exceeds the rating of the vertical deflection systems. This is a matter of an amplifier design and does not mean that the tube affords greater sensitivity along the horizontal dimension than along the vertical dimension.

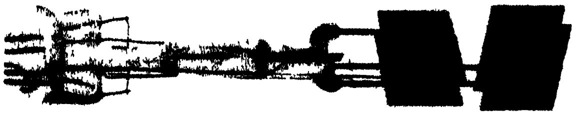
Referring to the improvement in deflection factor as a result of the amplifiers, it is not always the same in all equipment. The frequency range of the amplifiers has a great bearing on the amplification it affords. We shall not say much about this at this time, because the subject is treated separately in Chapter 10. However, we might mention that sometimes the deflection factor of the vertical deflection system in complete equipment covers a range of values, becoming greater (less sensitive), as the signal frequency is increased.

Commercial Electrostatic Deflection Systems

Several illustrations in Chapter 5 showed the detail of modern electrostatic deflection systems as a part of the mount which supported the electron gun. Here are several photographic views of the early and modern arrangements.

Standardization among electrostatic deflection systems is very widespread, although a few minor differences will be found. They are, as a rule, differences in physical dimensions and separation between the electrodes. Anyway very few radical changes have been made since the very first commercial cathode-ray tube. The most important has been a modification in the shape of the individual deflection plates.

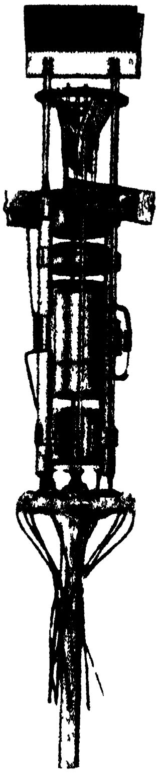
In the very early version of the tube, the deflection plates were straight surfaces parallel to each other throughout, as shown in Fig. 6-11. This is the deflection system used in the gas-focused Western Electric 224A described in Chapter 5. Such construction had one prime disadvantage in the light of the modern-day tube. If the separation between the surfaces of each pair of plates was kept to a minimum so as to intensify the field, thereby attaining a high order of sensitivity and a reasonable deflection on the screen, it tended to restrict the magnitude of the deflection voltage. If the deflection voltage exceeded a comparatively low figure, the path of the deflected beam would be bent sufficiently to strike the surface of the plate, thus impeding the advance of the beam to the screen. To overcome this by increasing the separation between the plates was undesirable, because it reduced the sensitivity. Thus the dimension of the deflection on the screen was limited, although the original design allowed more than ample deflection on the 4-inch screen which was used.



Courtesy Bell Telephone Labs.

Fig. 6-11.—Gas-focused Western Electric Type 224A.

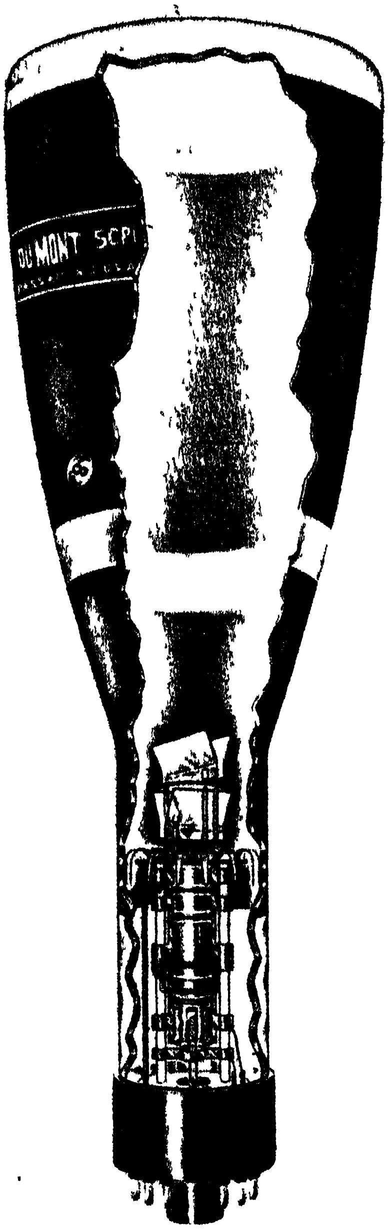
Modern deflection plates have been reshaped so as to follow the contour of the deflected beam, thus allowing increased deflection angles and larger traces without impeding the progress of the beam to the screen. Such plates can be seen in Fig. 6-12 and Fig. 6-13. They are parallel for a portion of their length, then they diverge. Of course, if excessive values of deflection voltage are applied to plates of this shape, the beam will be bent sufficiently and it will strike the plates. However, the modern design affords both increased sensitivity and greater latitude in the range of applicable deflection voltages.



Courtesy Du Mont Labs.

Fig. 6-12.—Type 3GP-A electron gun; note the shaping of the deflection plates.

The normal location of the conventional electrostatic-deflection system inside the neck of a modern cathode-ray tube is shown in Fig. 6-13. The relative locations of these plates with respect to the gun and the remaining portions of the tube can be seen clearly. The V-shaped structure immediately above the spider, which joins the accelerating anode in the gun to the second-anode coating on the inside of the tube, is the vertical deflection system. Above it is the horizontal deflection system. It is interesting to note that, considering the small dimensions of the beam, the active portion of each of the deflection plates is a very small part of the entire surface, although the entire area of each plate is a charged surface. The importance of properly orienting the opening of the upper set of plates relative to the deflected beam issuing from the lower set is readily evident. Improper positioning will not only place the beam in a nonuniform field, but may actually cause the plate to interfere with the passage of a portion of the beam, or at least cause it to strike the



Courtesy Du Mont Labs

Fig. 6-13.—Cut-away view of electrostatic type 5CP1.

nearest deflection plate sooner than the other plate, thus producing a nonlinear deflection on the screen for a linear deflection voltage.

Two-Dimensional Deflection

The process of two-dimensional electrostatic deflection is nothing more than a combination of two individual deflection processes. The description we shall give at this time will be very limited because several chapters later on are devoted to a full explanation.

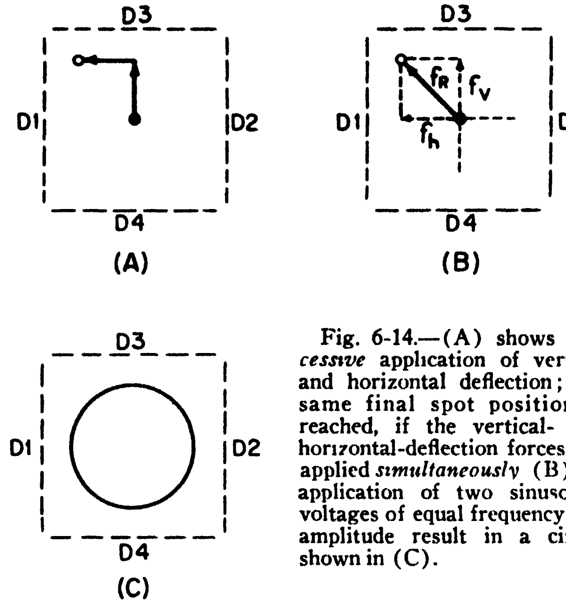


Fig. 6-14.—(A) shows *successive* application of vertical and horizontal deflection; the same final spot position is reached, if the vertical- and horizontal-deflection forces are applied *simultaneously* (B) the application of two sinusoidal voltages of equal frequency and amplitude result in a circle, shown in (C).

Let us consider the exaggerated illustration of deflection shown in Fig. 6-14A. Here we work with d-c voltages and we shall assume an appreciable lapse of time between the deflections in the two directions. The heavy dot in the illustration is the undeflected beam traveling toward the viewer out of the paper. The voltage applied to the vertical deflection plates *D*3-*D*4 applies a force f_v to the beam and it is moved upward as shown by the arrow. After reaching the upward limit of deflection, the voltage is applied to the horizontal deflection plates *D*1-*D*2, shown by dotted lines, and because of the polarity of this voltage, the beam is moved toward the left, so that the final position of the deflected beam is that shown by the circle. Now, if we assume *simultaneous* deflection forces of like magnitude, the same final location of the beam is obtained, as shown by the simple vector diagram in Fig. 6-14B. Here the vertical deflection force is f_v and the horizontal deflection force is f_h, the two acting at right angles to each other. The resultant force is established by the parallelogram method, the final location of the beam being indicated by the position of the circle. It is identical to that indicated before. Since the electron beam is made up of discrete particles, it stands to reason that it cannot be in two places at one time, no matter how great its velocity. Therefore, any position of the beam due to the application of two simultaneously acting forces is the resultant of the two. No matter what the character of the trace which appears on the screen, it is due to changing positions of the beam, that is, the motion of a single spot.

Just as the deflection of the beam due to a single voltage is determined by the magnitude of the voltage, the resultant deflection in a dual-deflection system is determined by the direction and magnitude of the two individual deflection forces. Thus in Fig. 6-14B, the final location of the beam corresponds to an angle of 135° because the two force vectors f_v and f_h are due to two voltages of equal magnitude and are positioned at 90° and 180° with respect to the positive X-axis.

If we apply two sinusoidal voltages of like frequency and magnitude and 90° apart, to the two deflection systems of Fig. 6-14B, the changing deflection forces applied to the beam will cause it to describe a circle, as shown in Fig. 6-14C. The actual development of this motion is too long to be described at this time, especially since a complete chapter is devoted to the subject later in this book. For the present, accept this statement as well as the fact that two such voltages produce a rotating voltage vector of constant amplitude and constant angular velocity. The how and why of this will be explained fully in Chapter 12.

Radial Deflection (J Scope)

A modification of the basic dual electrostatic-deflection system shown so far in this chapter is the tube wherein a third electrostatic deflection element is used. This type of tube is an outcome of the need for presentation along polar co-ordinates, although such presentation is possible without recourse to this third electrode. The Armed Forces made use of this type of tube in a number of radar units and the display was identified as the J-scan (tube type — 3DP1). Fundamentally, the tube is a standard electrostatic-type cathode-ray tube, except that an additional electrode in the form of a thin metallic rod is mounted at the center of the screen. The rod extends through the space between the screen and the deflection area, although not into the area occupied by the deflection plates. The other end of the rod protrudes through the screen where a snap-on clip enables electrical contact with the rod. The tube is shown in Fig. 6-15. The radial deflection electrode is clearly visible at the center of the screen. Some tubes, such as the 3CP1, also provide an etched scale on the screen for the determination of distances to the target, picked up by the equipment. Fig. 6-16 affords a general idea of the construction and the relative location of the radial deflection electrode and the conventional deflection plates.

Fig. 6-15.—Type 3DP1 J tube, showing the central deflection electrode protruding through the screen.

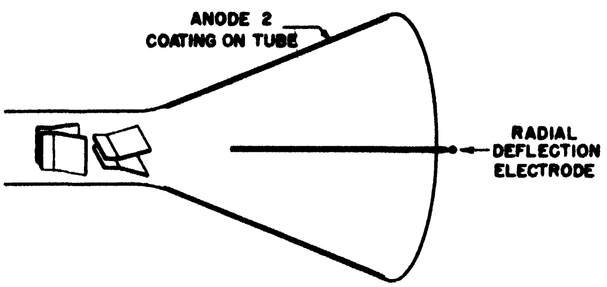


Fig. 6-16.—Internal construction of the J scope.

In the operation of this tube, the trace on the screen is organized along polar co-ordinates. Two sinusoidal voltages of equal magnitude and 90° out of phase are applied to the two pairs of deflection plates. The resultant trace on the screen is a circle, as in Fig. 6-14C. This is the time base for the display. The speed with which the spot travels around the screen describing the circle, or the angular velocity of the rotating voltage vector, is a function of the frequency of the voltages applied to the deflection plates. Thus the time base can be calibrated in terms of time in microseconds. Now, if a voltage is applied to the radial deflection electrode, it will set up a radial field between it and the second-anode *coating* on the inside of the tube envelope. Therefore, the beam which follows a circular path in the space between the deflection area and the screen is *moving through the electrostatic field* between the radial deflection electrode and the second-anode coating. How this second electrostatic field will affect the beam depends upon the instantaneous magnitude and polarity of the second field. By suitable arrangement, the signal voltage applied to the radial electrode can be such as to momentarily deflect the beam away from its path, that is, kick the beam outward, causing an instantaneous increase in the length of the radial voltage vector which is describing the circle. The extent of this "kick" is a function of the amplitude and time duration of the radial deflection voltage.

Thus in Fig. 6-17A, we show the circular time base without any voltage applied to the radial deflection electrode. The circular trace is the result of the two sinusoidal voltages applied to the two pairs of deflection plates. Now a momentary signal voltage in the form of a pulse is applied to the radial deflection rod, and this pulse voltage has such polarity as to kick the beam outward. The appearance of the trace now is as

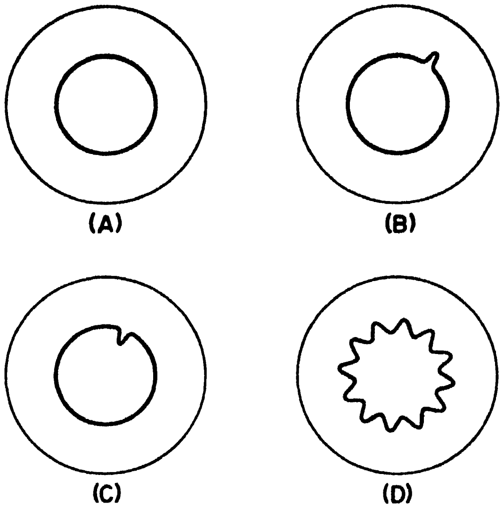


Fig. 6-17.—(A) shows the circular time base produced on the screen of the J scope before a deflection voltage has been applied. In (B), the circular base has been modified by a momentary voltage pulse applied to the radial deflection electrode, in (C) a pulse of the opposite polarity is applied to the electrode, and (D) shows the appearance of the trace after a sine-wave voltage is impressed on the radial deflection electrode.

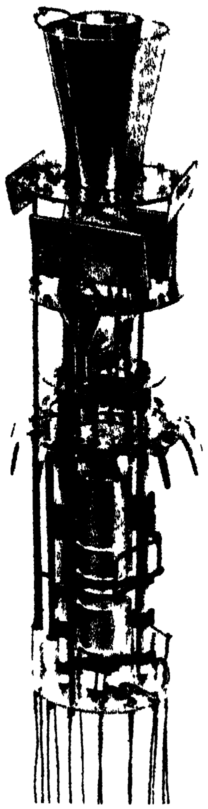
in Fig. 6-17B. A change in the polarity of the radial deflection pulse can cause the trace to kick inward as in Fig. 6-17C. If the voltage applied to the radial deflection electrode is a sine wave, the final trace on the screen will appear like that shown in Fig. 6-17D.

It is significant to note that the frequency of the signal which produces the time base (circle) is not changed when the radial deflection voltage is applied; in other words, the voltage applied to the radial deflection rod does not influence the operation of the regular deflection system. The circular moving beam is changed in its orbit by the radial deflection field *after* it has left the regular deflection system.

The latter point is important because of circuit modifications which enable the development of a similar trace without recourse to a second radial deflection field. This is explained fully later; for the present, let it suffice to say that the voltages applied to the regular deflection plates, which develop the circular time base, are so modified momentarily by a supplementary signal voltage, as to produce traces like those shown in Figs. 6-17B through D. Without this supplementary voltage the trace appears like Fig. 6-17A.

Radial Deflection (Truncated Cone System)

A type of tube which affords radial deflection without recourse to the external radial deflection electrode is shown in Fig. 6-18.[1] This tube makes use of the usual electron gun, the conventional dual set of parallel deflection plates, and has in addition two truncated cones mounted in the space between the deflection system and the screen of the tube. These are located above the uppermost set of deflection plates, one cone being concentric with the other, with a space between them.



Courtesy Electronic Tube Corp.

Fig. 6-18.—Gun mount for the ET 3000 radial deflection cathode-ray tube.

The application of two voltages equal in amplitude and frequency 90° apart in phase to the deflection plates creates the circularly moving beam. By suitable values of these voltages the diameter of the circle traced by the beam is such as to cause it to rotate within the space between the two cones. The signal voltage required for the radial deflection of the beam, which may be described as modulation of the beam, is applied across the two cones, thus developing a field between these surfaces. With alternating voltages applied to the radial deflection system, the modulation of the beam will take place in both directions and the trace developed will resemble that shown in Fig. 6-17D. If a unidirectional voltage is applied across the cones, the beam will be deflected outward or inward, depending upon the instantaneous polarity of the voltage.

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[1] von Ardenne, Manfred, "A new polar co-ordinate cathode-ray oscillograph with extremely linear time scale," *Wireless Eng.*, vol. 14, pp. 5-12, January, 1937.

The physical space between the cones determines the maximum modulation, therefore, the maximum amplitude of the voltage applicable for radial deflection. Excessive beam modulation voltage will cause the beam to strike either or both inner-cone surfaces, thus interrupting the advance of the beam to the screen. According to the manufacturer, the design of these systems permits a peak-to-peak modulation of about ¾ inch. Two screen diameters are available, 3 inches and 5 inches. The latter affords circular time bases which may be varied from 1⅞ inches to 3¼ inches in diameter with the aforementioned amount of radial deflection, without interference from the cone structure. This, of course, assumes that the recommended second-anode and post-deflection accelerating voltages are employed. Comparative tests made by the manufacturer indicate that the radial deflection sensitivity of this type of construction is about six times that available with the 3DP1.

Radial Deflection Factor

It is understandable that the radial deflection system also will bear some rating of the magnitude of voltage required to produce a certain value of radial deflection. In other words, radial deflection systems also have deflection-factor ratings. For the tube shown in Fig. 6-18, the rating is 90 volts d.c. per inch.

Combined Radial and Rectangular Deflection Tubes

By combining in one tube envelope systems for rectangular and radial deflection, both types of traces may be produced. An example of this is shown in Fig. 6-19, wherein three individual guns are employed. Two of these are associated with conventional deflection plates for rectangular co-ordinate traces, whereas the third is intended for polar co-ordinate traces. It is interesting to note that tubes specifically designed for radial deflection, such as the one which utilizes the radial deflection electrode, and those which use the cones, are not very satisfactory for the production of traces organized along rectangular co-ordinates. The reason is the presence of the radial deflection elements in the path of the beam. This is truer in the case of the cone structure than in the case of the single-rod type electrode, but in either case, it is a limitation.

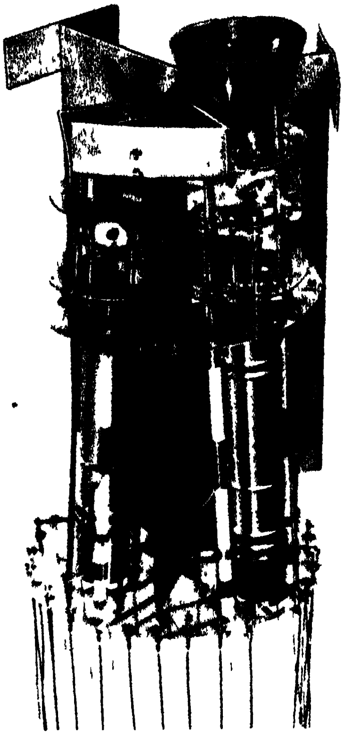


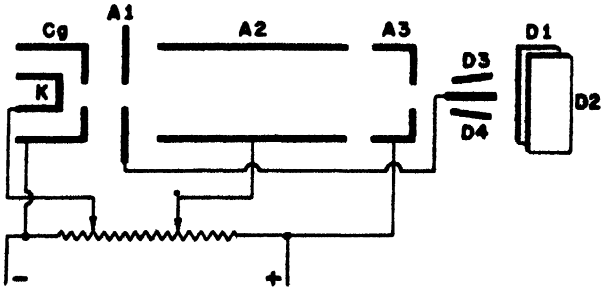
Fig. 6-19.—Internal construction of the Type 73HU 3-gun cathode-ray tube.

Courtesy Electronic Tube Corp

Such restriction in use is absent in the system which enables polar co-ordinate presentation with the conventional tube, but utilizes specially designed amplifiers. It is, therefore, evident that the selection of one type of tube or the other is a matter of what is to be shown, as well as the consideration of all of the factors involved.

The Split-Beam Deflection System

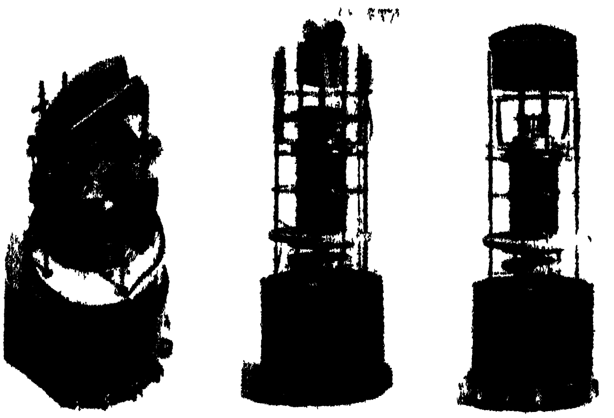
Modification of the usual deflection system used for rectangular co-ordinate traces is to be found in a tube introduced in the United States from a British-made oscilloscope. This is the Cossor split-beam tube, whereby two separate traces may be made to appear on the tube face. This differs in several ways from the usual dual beam using two electron guns. First, a single electron gun is used, and second, the beam is split into two parts, each separately controllable after leaving the electron gun.



After Television and Short-Wave World

Fig. 6-20.—Schematic of electrode system of Cossor split-beam cathode-ray tube.

The tube made its appearance as early as 1939,[2] but only after the war did complete equipment utilizing it become available in the United States. The full details of the complete oscilloscope are to be found in Chapter 22; in the meantime, the schematic representation of the gun and deflection-plate assembly are shown in Fig. 6-20. Constructional details are shown in Figs. 6-21A, 6-21B, and 6-21C. The cathode lens is formed between the control grid and the accelerating anode *A*1. The principal focusing lens is formed between anodes *A*2 and *A*3. These designations differ from those used in this country, but conform with the descriptions as given by the author.

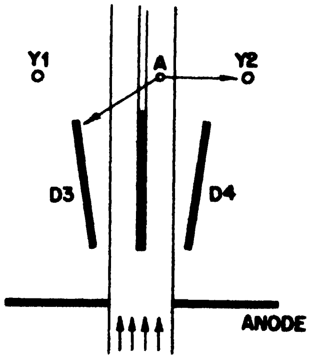

(A) (B) (C)

Courtesy Cossor Ltd

Fig. 6-21.—Constructional details of the British Cossor split-beam cathode-ray tube. The splitter electrode is visible set between two deflection plates below the top of (C).

Close to *A*3 and connected to it, and directly in the path of the beam issuing through the aperture of *A*3, is a thin piece of metal, the splitter plate. This divides the beam in half and acts as electrostatic shield between the two vertical deflection plates *D*3 and *D*4, so that a potential applied to one of the plates will act only on one half of the beam. The splitter plate is kept short and the deflection plates are relatively close together so as to attain the minimum loss in sensitivity. In a given design of cathode-ray tube, if the ratio of deflector-plate separation and deflector-plate length are kept constant, the sensitivity will remain practically constant no matter how short the plates. Of course, if the separation between the plates is made too small, the permissible deflection voltage likewise will be kept to an extremely small value in order to prevent interference with the travel of the beam toward the screen. Owing to the limited size of the splitter plate, the shielding of the two halves of the vertical deflection system is not perfect, and unless proper correction is applied, a deflection voltage applied to one vertical deflection plate would affect the other. Since the screening capabilities are limited consistent with minimum loss of efficiency, an artifice is used to attain the desired results. This is the use of two additional deflector plates in the form of wires. Their action is as follows. Examine Fig. 6-22. Suppose that a positive deflection voltage is applied to *D*3, then the half-beam nearest to it will be attracted toward it. However, electrons in the neighborhood of point *A*, also will be attracted toward the positively charged plate *D*3 as shown by the arrow, so that this half of the beam also will be deflected somewhat to the left. In order to keep this latter half of the beam stationary, two additional deflector plates, in the form of wires *Y*1 and *Y*2 are used. The former is connected to *D*4 and the latter to *D*3. Under the conditions described, the electrons at *A* will feel an attracting force issuing from *Y*2 as well as from *D*3. With voltages applied to both deflection plates, the attracting force due to the voltages on the other plate will be balanced out by the oppositely polarized deflection wires, so that unwanted deflection is prevented.

Fig. 6-22.—The diagram illustrates the action of the deflector plates in the Cossor split-beam tube.



After Television and Short-Wave World

The optical equivalent of such a beam-splitting system is shown in Fig. 6-23, and it is seen that the undeflected position of the beam is at the geometric center of the tube. It is only when deflection voltage is applied to either or both vertical deflection plates, that the separation of the beam into two halves becomes visible.

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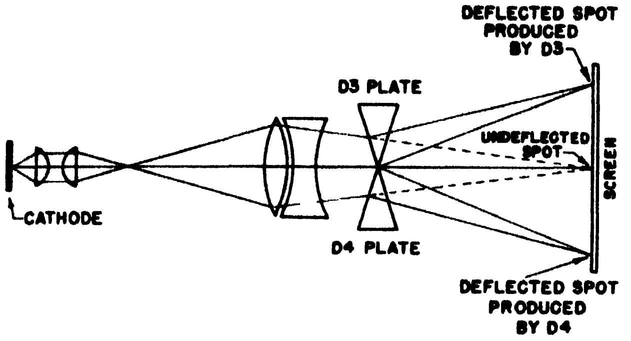
[2] Fleming-Williams, B. C., "The Double-Beam Cathode-Ray Oscillograph," *Television and Short-Wave World*, vol. 12, p. 457, August, 1939.



After Television and Short-Wave World

Fig. 6-23.—Optical analogy to the action of the electron gun in the Cossor split-beam cathode-ray tube.

Electrical Characteristics of Electrostatic Deflection Systems

Some of the characteristics already have been referred to, as for example, the deflection factor and sensitivity constants. There are, however, several more which deserve comment, since they not only appear in the tube specifications, but have an influence on the utility of the systems.

Capacitance of Deflection System

Since in effect each pair of deflection plates forms an electrical capacitor, it is natural that there be associated with each pair the constant of electrical capacitance. Inasmuch as the design of all deflection systems of this type is not the same, the electrical capacitance presented by the systems in various tubes will be different. They may vary from a few micromicrofarads per pair of plates to perhaps 12 to 15 $\mu\mu f$.

When used for a-c work, each pair of plates, therefore, offers a certain amount of impedance across whatever may be the circuit connected to the deflection plates. When the plates are connected directly to a source whose output is to be investigated, the impedance of the deflection system comprises a load on the source. Therefore, the higher the impedance of the deflection system the less is this load; in turn, the impedance at any one frequency is highest, when the capacitance of the deflection plates is a minimum. It might be well to mention that the capacitance introduced by the cables, which connect the deflection plates to the source under investigation must be taken into account, and that this capacitance should be considered as being in parallel with the deflection plates to which the leads are joined.

Deflection Defocusing

It would normally be expected that the focused beam leaving the electron gun would remain focused while being deflected and arrive at the screen as a sharply defined image. Such, however, is not always the case. The reasons are very interesting. While it may not have been made evident during the preceding discussion of the electron beam, the beam which enters the deflection area is not as narrow as when it arrives at the screen, since the focal point of the principal focusing lens is the screen, rather than the deflection area. Consequently, the beam passing through the deflection-plate fields has a relatively substantial cross section. So much so that the beam can be changed from a circle to an ellipse due to the increased velocity of the electrons on that side of the beam which is closer to the positively charged plate than the electrons on the other side of the beam. The closer the beam approaches to the positively charged plate the more aggravated becomes this situation, giving rise to defocusing of the spot as it approaches the limits of the screen. Thus the spot changes from a circle to an ellipse with its major axis in the direction of the deflection.

Change In Sensitivity

Another effect is to be found in those deflection systems which employ grounded plates. Some tubes were and still are made with one plate of the vertical pair and one plate of the horizontal pair joined to the accelerating anode. It has been established that the velocity of the beam through the deflection area is related to the sensitivity of the tube. If for any reason during the development of a trace of alternating character, a change in tube sensitivity takes place due to changing behavior of the tube, the trace becomes distorted.

Such an effect comes about in the tubes mentioned as the result of a change in the velocity of the beam, and an associated change in the sensitivity. If one plate of a pair is joined to the accelerating anode, the plate assumes the potential of the anode. Thus if the second-anode voltage is 1,000 volts and an a-c deflection voltage of 100 volts is applied across the deflection plates, the voltage of the "free plate" goes to 1,100 volts on the positive half-cycle of the deflection voltage and to 900 volts on the negative half-cycle with respect to the cathode. Thus on the positive half-cycle the beam is deflected upward into the higher-voltage region and acquires an increased velocity and, consequently, lower sensitivity. During the negative half-cycle of the deflection voltage, on the other hand, the beam is pushed downward into the lower-voltage and lower-velocity region with an attendant increase in sensitivity.

The cure for this is the use of balanced deflection systems, where neither of the deflection plates in the vertical or horizontal pair are joined to the anode; a

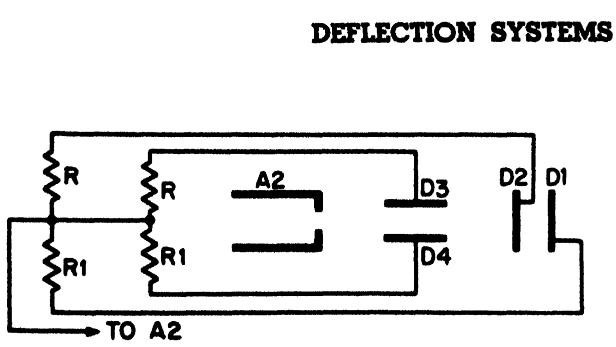
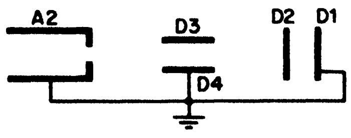


Fig. 6-24.—A balanced deflection system, where neither of the deflection plates are joined directly to the accelerating anode.

balanced circuit may be arranged, as shown in Fig. 6-24. This is only attainable when each deflection plate is equipped with its own individual connecting terminal. The values of R and $R1$ are both equal, their specific value depending on manufacturer's recommendations. Such a system is known as a *balanced* deflection system in contrast to the *unbalanced* system shown in Fig. 6-25. The balanced system is fed by push-pull amplifier circuits, which are discussed in Chapter 10.

Fig. 6-25.—Unbalanced deflection system; one plate of each deflection pair is connected to the accelerating anode *A2.*



In addition to a change in sensitivity with an unbalanced deflection system, an accompanying effect of the above condition is a change in focusing; this is related to the effect described under "Deflection Defocusing".

Still another effect of asymmetrical connection of the deflection plates is known as "trapezium" distortion. This is the development of a trapezoidal pattern instead of a square or a rectangle pattern, when equal amplitude voltages (appreciably different in frequency) are applied to the two deflection systems. This condition also is known as *area distortion.*[3] As to the reasons for it, they are twofold, one being the absence of a constant mean potential across the deflection system relative to the anode voltage, and the other being the fact that the vertical deflection is reduced during the time that the free plate in the horizontal deflection system is positive. Both of these related conditions may be eliminated by the use of a symmetrical deflection system (push-pull input). Sometimes trapezium distortion is due to misalignment of the deflection systems relative to the axis of the tube. Several methods for correcting trapezium distortion have been suggested,[4] which revolve around shaping of the plates. These are suitable for both symmetrical and asymmetrical, or push-pull and single-ended input circuits. Tubes manufactured in the United States do not employ these special shapes, possibly because of the popularity of push-pull circuits, which are relatively free of this problem.

Phase Shift Between Deflection Fields at UHF

This phenomenon is native to electrostatic tubes, but is important only in connection with very high frequencies.[5] It can be disregarded completely in the usual run of cathode-ray-tube equipment, intended for maintenance operations, and even for research, providing that the frequencies applied to the vertical and horizontal plates do not exceed about 10 Mc, and that the second-anode voltage approximates about 2,500 volts.

At higher frequencies and lower anode voltages, the problem is not too severe if the ratio of frequencies applied to the vertical and horizontal plates varies by at least 4 to 1. It is to be noted that the action to be discussed is associated only with the tube and does not include the amplifying equipment. The latter is an entirely different matter, discussed later in this book.

We remarked previously that due to the physical arrangement of the deflection systems along the path of the electron beam, both fields do not act simultaneously on the beam. The deflection system closest to anode 2 (vertical) acts on the beam first, then, sometime later, when the beam arrives in the region of the horizontal deflection plates, that system performs its deflection function. Therefore, a certain period of time does elapse before the electrons in the beam complete their journey from the center of the field of the vertical deflection plates to the center of the field of the horizontal deflection plates.

In Fig. 6-26 is a representation of the two deflection systems relative to the path of the electrons, and shows what is meant by the distance between the midpoints of the two deflection fields. For the sake of discussion, let us assume that the facing edges of the two sets of plates are 0.125 inch apart; also assume that the length of each deflection plate is 0.875 inch, so that the center-to-center distance is 1 inch or 2.54 centimeters.

This hypothetical tube is operated at an accelerating voltage of 1,000 volts, and according to previous cal-

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[3] Puckle, O. S., "Electrostatic deflection," *Wireless World,* vol. 42, p. 375, April 28, 1938.

[4] Fleming-Williams, B. C., "Trapezium distortion in cathode-ray tubes," *Wireless Eng.,* vol. 17, p. 61, February, 1940.

[5] Hollman, E., "Hochfreq. und Electroak," vol. 54, 1939.

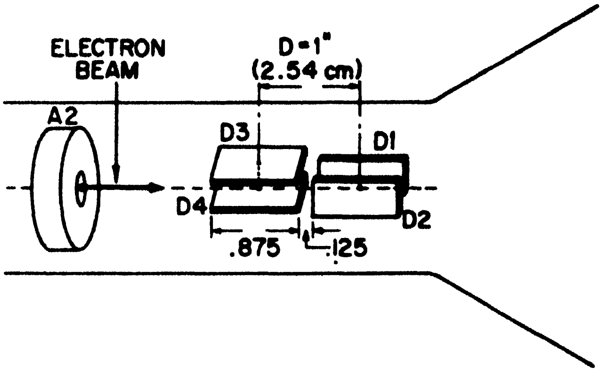

Fig. 6-26.—Illustration shows the distances involved in the two deflection systems; *D* is the distance from the center of one deflection system to the center of the other system.

culations, the moving beam will span the distance of 1 inch in 1352 micromicroseconds. An electron leaving the center of the *D*3-*D*4 field will not arrive at the center of the *D*1-*D*2 field until 1,352 micromicroseconds later. If a signal is applied simultaneously to both sets of plates, it will be in phase on both plates relative to the source. As far as the electron beam is concerned, however, it will not feel the effects of these two signal voltages at the same instant. The *D*1-*D*2 field cannot act on the beam until 1,352 micromicroseconds after the *D*3-*D*4 field has started to act on the beam. At higher accelerating voltages, this time lag will be less, and at lower accelerating voltages, it will be greater.

Whether the time lag is or is not important depends on the frequency of the signal applied to the two deflection systems and the desired accuracy of the work being done. Suppose that the applied signal is 20 Mc. What will be the angular distance traveled by the wave applied to the *D*1-*D*2 plates during the 1,352 micromicrosecond transit time of the electron beam as it travels from *D*3-*D*4 to *D*1-*D*2?

The angular distance traveled by a wave in a time t is

$$2\pi ft \text{ radians} \qquad (6\text{-}6)$$

and expressed in degrees it is

$$2\pi ft \times \frac{360}{2\pi} = ft \times 360. \qquad (6\text{-}7)$$

Substituting the quantities stated

$2\pi = 6.28$

$f = 20 \times 10^6$ cps

$t = 1.35 \times 10^{-9}$ seconds

and solving the above, the voltage applied to *D*1-*D*2 will have passed through 9.7° during the time that the electron beam traversed the space between the midpoints of the two fields. Thus a phase shift of 9.7° exists between the traces, because of the transit time of the electron beam.

Increasing the accelerating voltage and keeping the signal frequency constant would reduce the phase shift, while keeping the accelerating voltage constant and raising the frequency would increase the phase shift. If the frequency is raised by such an amount that the time taken to complete one cycle is equal to the transit time of the beam between the two deflection pairs, then the phase shift will be 360°, which is equivalent to 0°. The same situation would occur, of course, when the transit time is equal to any multiple of the periodic time of the voltage; i.e., when

$$\text{transit time} = \frac{n}{f} \qquad (6\text{-}8)$$

where n is any integer.

For the transit time of 1.35×10^{-9} seconds in our present example, the frequencies for which no phase shift occurs could, therefore, be any multiple of $\frac{1}{1.35 \times 10^{-9}}$ or 741×10^6 cps (i.e., 741 Mc). At multiples of one-half of this frequency, i.e., 370.5 Mc, the phase shift would be 180°, which effectively turns the waveshape upside down on the screen.

The phase difference in degrees between any two signal voltages applied to the deflection plates relative to the transit time of the electron beam may also be determined from the equation

$$\frac{D1 \times f}{V} \times 360° \qquad (6\text{-}9)$$

where

$D1$ = the distance between the centers of the deflection system in centimeters

f = the frequency of the applied signal

V = the velocity of the beam in centimeters per second at the accelerating voltage being used.

In the example cited, replacing the symbols with the operating values results in a phase shift of

$$\frac{2.54 \times 2 \times 10^7}{1.875 \times 10^9} \times 360 = 9.7°.$$

For ordinary applications of the cathode-ray oscilloscope, this is not important. It is pertinent to phase measurements rather than to waveform observations and analysis, and even then only when very high frequencies are involved. For this type of work, special oscilloscopes have been designed, wherein the beam velocity is extremely high due to the application of high voltages. In the oscilloscope mentioned above, the accelerating voltage is 50,000 volts.

A great many applications of the oscilloscope involve a lower frequency on the horizontal plates than on the vertical plates, such as when waveform measurements are made, or when two different frequencies are being compared. Both are described in detail later; in the meantime, it should be understood that when the above-mentioned condition prevails, the phase shift is reduced in proportion to the ratio between the two signal frequencies, assuming that the beam velocity is held constant. Thus in the example cited, if the frequency applied to the vertical plates is 20 Mc and the frequency applied to the horizontal plates is 5 Mc, the phase shift would be

$$9.7 \times \frac{5}{20} = 2.4°.$$

If the conditions were reversed, so that the angular velocity of the signal applied to the horizontal plates was greater than that of the signal applied to the vertical plates, the phase shift would be increased in the proportion to the ratio between the two signal frequencies.

Beam Transit Time and Deflection Sensitivity

Another aspect of the beam transit time is the relation between the time required for the beam to traverse a *single deflection field* and the frequency of the field. This effect was not evident when we established the deflection due to d-c voltages applied to the deflection plates. Neither is it important when the frequency of the deflection voltage is 100 Mc or *less,* and the accelerating voltage is *at least* 1,000 volts. Actually, the effect becomes less important as the accelerating voltage is increased because the transit time is decreased. This means that the average user of a cathode-ray oscilloscope for maintenance operations, in fact, even for research operations, need not concern himself with the previously mentioned relationship unless operation at very high frequencies or in the microwave region is carried on.[6]

Essentially the problem is that, if the beam transit time through the field corresponds to the periodic time of one or more whole cycles of the deflection voltage, the over-all deflection sensitivity is reduced to zero. This is shown in Fig. 6-27 for a single cycle, and applies equally well to any whole number multiple. The signal is viewed in terms of the action of the upper plate and it is assumed that the phase of the signal is such that it is of zero magnitude at the instant the beam enters the field. Then the beam will be deflected upward during the time corresponding to one half-cycle and downward during the time equal to the other half-cycle. Since the magnitude of deflection is the same in both directions, the net result is zero deflection when the beam leaves the field, since the deflection in one direction is later offset by deflection in the other direction.

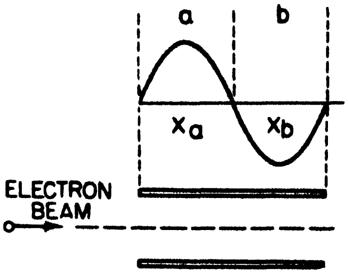

Fig. 6-27. — Here the beam transit time for one pair of deflection plates corresponds to one whole cycle of the periodic time resulting in zero effective deflection.

For other periodic times of the deflection voltage, and for other transit times, the deflection sensitivity will vary between some finite value and zero; also it is always less than the values which would be obtained with a d-c voltage applied to the deflection plates. The relative deflection sensitivity is expressed by

$$\frac{\sin\left(\frac{\omega t}{2} \times 57.3°\right)}{\frac{\omega t}{2}} \qquad (6\text{-}10)$$

where

$\omega = 2\pi f$

t = the beam transit time in seconds.

A graph which shows the effect of transit time on the deflection sensitivity is given in Fig. 6-28. To show the application of this chart, let us assume that the accelerating voltage is 1,000 volts and the length of the deflection field (plates) is 2 cm. With this value of voltage on anode 2, the time required for the beam to traverse 2 cm is 1,065 micromicroseconds. Suppose that the frequency is 50 Mc. Then

$$\frac{\omega t}{2} = \frac{6.28 \times 50 \times 10^6 \times 1.065 \times 10^{-9}}{2} = 0.167$$

and the relative dynamic sensitivity according to Fig. 6-28 is almost 100 per cent or equal to d.c. Now let us imagine that the frequency is 500 Mc and the transit time is the same. Then

$$\frac{\omega t}{2} = 1.67$$

which corresponds to a relative deflection sensitivity of approximately 75 per cent. Thus a correction factor of 1.33 would be necessary to bring the deflection sensitivity to that obtained with d.c.

As a whole this condition may be ignored when operating with sine waves and frequencies below the values stated, but if the transient response is to be investigated, even though the fundamental frequency may be low, the harmonics in the wave may be of sufficiently high order for a loss in deflection sensitivity to occur.

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[6] Libby, Lester, "Cathode-rays for uhf," *Electronics,* vol. 9, pp. 15-17, September, 1936.

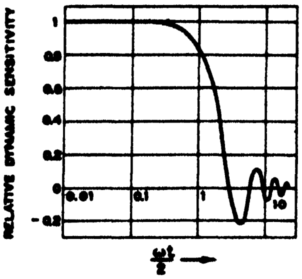

Fig. 6-28.—The lowering of the dynamic sensitivity due to electron transit time.

Courtesy Proc. I.R.E.

In that event, the trace on the screen will not be a true representation of the deflection voltage applied. A three-beam oscillograph for recording at frequencies up to 10,000 Mc is described by Lee.[7]

ELECTROMAGNETIC DEFLECTION SYSTEMS

Magnetic deflection systems see very little use in research and maintenance oscilloscopes. Almost all such devices utilize electrostatic deflection, primarily because of the linear relationship between the deflection voltage and the deflection on the screen, and also because of the wide range of frequencies which it can accept. The frequency-impedance relationship of the deflection windings is a limiting factor, especially when nonsinusoidal voltages and currents are involved. The special current shape required for linear deflection by a magnetic deflection system introduces numerous restrictive conditions. These conditions are inimical to the utility of the deflection system in research devices operating over a range of frequencies.

Magnetic deflection is commonplace in television systems and in equipment operating on the pulse-echo principle, of which Armed Forces and navigational radar systems are significant examples, and in many devices which function over a very limited frequency range. In both instances, the frequency range is fixed.

The principles underlying magnetic deflection were described in Chapter 3; therefore, they need not be repeated here. Other details associated with the system, not necessarily a vital part of the principles, were omitted from that discussion, and they will be treated in this chapter.

Like the electrostatic system, the magnetic system is a two-dimensional deflection arrangement, which fundamentally produces a trace organized along rectangular co-ordinates. This need not always be so, but it is the basic form of display. A basic physical arrangement of the coils is shown in Fig. 6-29A. Two pairs of coils are located around the neck of a cathode-ray tube, each pair comprising a complete deflection unit. One pair is responsible for the vertical deflection, and the other pair develops the horizontal deflection. Either pair may be used for either deflection, unless special design considerations dictate otherwise. Schematically, the systems are shown in Fig. 6-29B. Coils *A-A'* are connected in series, and so are coils *B-B'*.

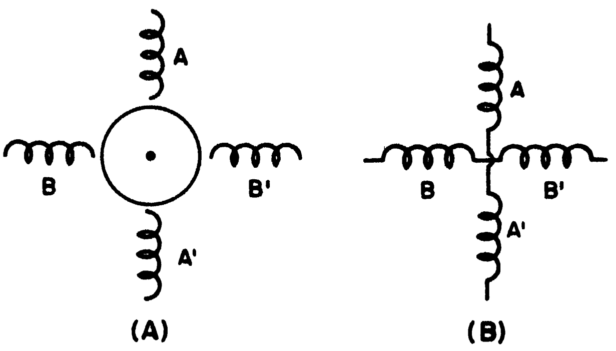

Fig. 6-29.—Physical arrangements of deflection coils. (A) shows the positioning of the two sets of coils at right angles around the neck of the tube, and (B) the series connection of each pair of coils.

While the axes of the two systems cross each other inside the tube, at a point along the path of the undeflected electron beam, they are not intended to indicate the direction of deflection nor the direction of the lines of flux of the fields created between these coils.

The series connection between the coils of each pair is by far the most popular arrangement found in electromagnetic deflection systems, especially in television receivers, but is nevertheless not the standard in every unit which employs magnetically deflected electron beams.

The magnetic fields created by the currents in such coils are illustrated in Figs. 6-30A, B, and C. Three different pictorial arrangements of deflection coils were deliberately selected in order to minimize the possibility of confusion due to the impression one might gain from a single arrangement such as the one shown in Fig. 6-29A, wherein the axes of the coils penetrate the field area. The physical orientation of the coils may be at least three kinds and still attain the necessary beam deflection action. For that matter, the coils themselves may be of varied construction.

For example, in Fig. 6-30A the coil structure is the same as shown in the fundamental illustration of Fig. 6-29A. The flux lines due to the current in *A-A'* are parallel to the axes of these coils. The same is true of coils *B-B'*, and their field. As to the two fields, an *ideal* condition is assumed, namely, that *every flux line of*

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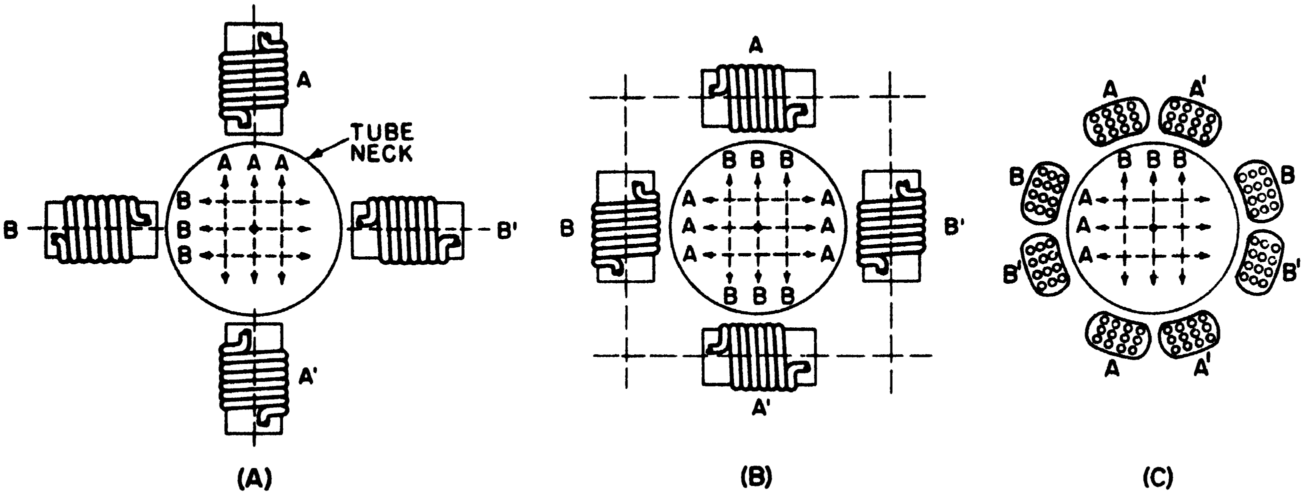
[7] Lee, Gordon M., "A three-beam oscillograph for recording at frequencies up to 10,000 Mc," *Proc. I.R.E.*, vol. 34, p. 121W., March 1946.



(A) (B) (C)

Fig. 6-30.—Magnetic fields created by various arrangements of deflection coils around the neck of the tube. (A) corresponds to the coil structure of Fig. 6-29A, in (B), the coils have been turned through 90°, and (C) shows the field produced by flat rectangular windings bunched into cylindrical bundles.

one field is perpendicular to every flux line of the other field, everywhere in the field.

In Fig. 6-30B, each of the coils has been turned through 90°. They still are solenoid windings, although not necessarily single-layer coils. The axes of the coils in a pair are parallel to each other, but are at right angles to the axes of the other pair, as indicated by the dotted lines. Though the axes do not cross the tube, the fields associated with these coils do penetrate the space in the tube throat. The lines of flux have been turned through 90° because the axes have been rotated through that angle. Since the axes are mutually perpendicular, the lines of flux of one field are perpendicular to the lines of flux of the other field.

A third arrangement of the windings is shown in Fig. 6-30C. Here coils *A* and *A′* are flat, rectangular windings with the turns bunched into a cylindrical bundle. The same is true of coils *B* and *B′*. Each coil has two long sides and two short sides. The circles in the illustration are cross-section views of the turns in the long sides. After proper forming, each coil is bent into a semicircular shape and positioned along the tube in such manner that the long sides run parallel to the neck of the tube and hence to the undeflected beam path. One long side of coil *A* and one long side of coil *A′* are positioned adjacent to each other on one side of the tube envelope and the other two long sides are positioned adjacent to each other on the other side of the tube. (The shape of such a winding is shown in Fig. 6-35A, and the assembly of the complete deflection yoke is described later in this chapter.)

The fields due to these windings have the same direction as those created by the arrangement shown in Fig. 6-30B, because the plane of the long sides of the special windings is parallel to the plane of the corresponding cylindrical turns in the windings of Fig. 6-30B; hence the field lines will be perpendicular to the planes of each coil, respectively.

Direction of Magnetic Field versus Direction of Deflection

Unlike the electrostatic system wherein the direction of deflection conforms with the direction of the lines of force, the magnetic system deflects the electron beam in a direction at right angles to both the direction of the flux lines and the direction of advance of the charge. Accordingly, the *A-A′* field in Fig. 6-30C would cause a beam of deflection as indicated in Fig. 6-31A. For that matter, the same is true of the *A-A′* field shown in Fig. 6-30B. Although it has not been so shown, the axes of the windings in Figs. 6-30B and 6-30C are the same.

The direction of the beam deflection due to the *A-A′* field in Fig. 6-30A would be turned through 90° with respect to that of Fig. 6-30B and C, since, as we stated earlier, the *A-A′* fields as well as the *B-B′* fields of Figs. 6-30B and C have also been turned through 90°. This is shown in Fig. 6-31B.

We can summarize all of this by saying that when the axis of a magnetic deflection system is perpendicular to the undeflected path of the electron, the deflection of the electron will be at right angles to the axis of said deflection system. This is evident when you compare Figs. 6-30A through C, with Figs. 6-31A and B. If the axis of the deflection coil is oriented horizon-

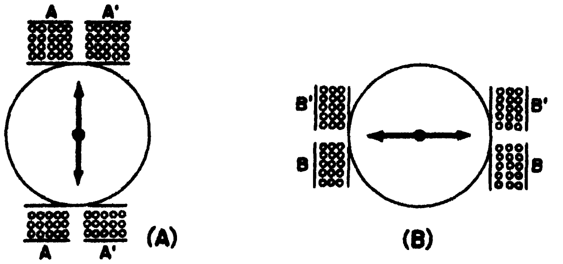


Fig. 6-31.—(A) shows the deflection caused by the field corresponding to Fig. 6-30C. The direction of deflection corresponding to the arrangement in Fig. 6-30A is shown in (B).

tally relative to the tube screen reference axis, then the beam deflection will be in the vertical or Y direction, and if the orientation of the coil is vertical relative to the horizontal reference axis, then the beam deflection will be in the horizontal or X direction.

The Composite Magnetic Field

Another point of interest relative to the two fields in Fig. 6-30 is the box-like appearance of the deflection system as a whole. Unlike the electrostatic arrangements, wherein the deflection plates are displaced laterally along the path of the electron beam, in the usual magnetic system, all four coils are equidistant from any one point along the path of the beam, so that both the vertical and horizontal deflection forces act simultaneously on the beam.

This may seem a bit confusing because it implies the presence of two magnetic fields within the same space, a condition previously stated as being impossible because the cross-over point of each line of force would have two directions. What actually happens is that a single composite field is established which produces the resultant deflection (as shown in Fig. 3-13 for different magnitudes of the individual fields.) At any one instant, the deflection is proportional to the vertical *and* the horizontal deflection forces.

Constants of Magnetic Deflection

Since the behavior of the magnetic deflection system is independent of the direction of the deflection, we can analyze the constants of such systems for a single pair of coils, although as we stated earlier in this chapter, two-dimensional deflection is necessary for proper portrayal of the majority of phenomena. It is only when the quantity involved is scalar, that is, has magnitude only, that one-dimensional deflection is sufficient.

In Fig. 6-32 is shown a pair of deflection coils positioned along the neck of a cathode-ray tube. The basic location of these coils, rather than the location of the magnetic field, is stipulated by the tube manufacturer on the outline drawing of the tube, which usually is a part of the tube specifications. We are referring to the point O along the tube throat identified as the ***effective center of deflection***. The location of this point in the case of the 10BP4 is shown in the outline drawing of Fig. 6-33.

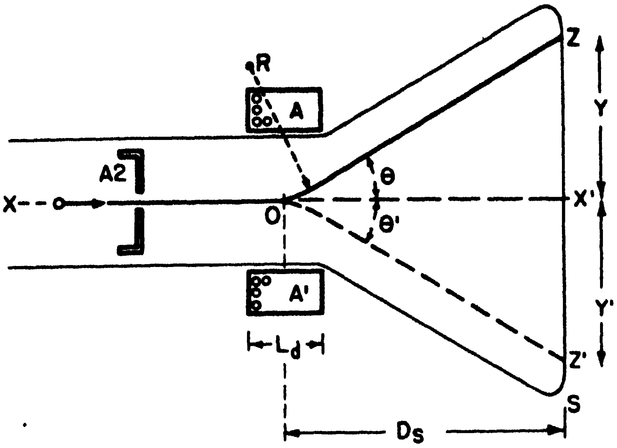


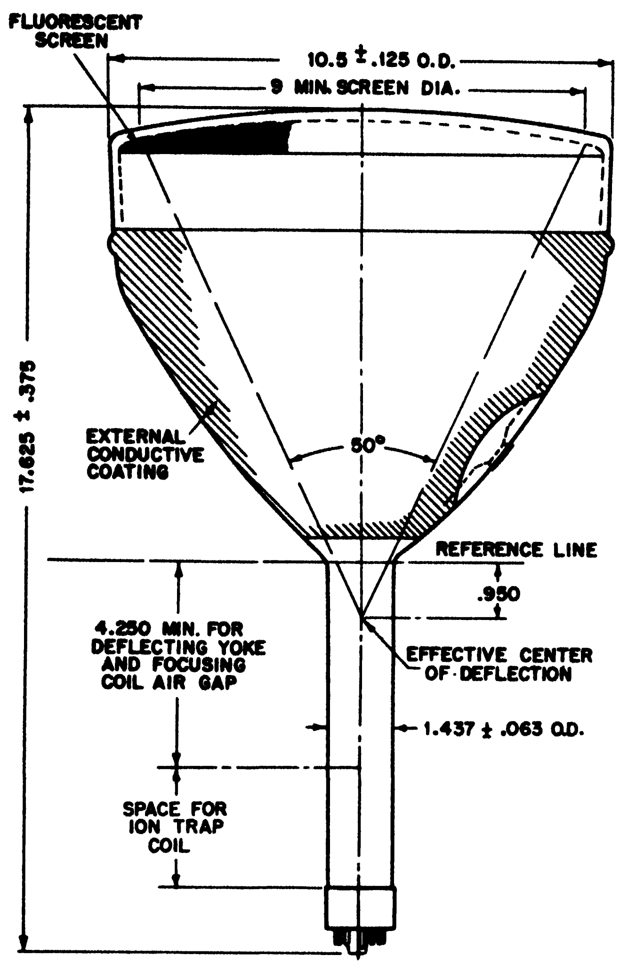
Fig. 6-32.—The proper positioning of the deflection coils with respect to the geometric center of deflection of the cathode-ray tube.

Referring again to Fig. 6-32, the effective center of deflection O is the imaginary point of origin of the electron beam for its final path of travel to the screen after the field has deflected it. This point is located at the midpoint of the field and should line up with the actual center of deflection, so that a deflection equal to the rated deflection angle θ of the tube will not experience any interference from the envelope of the tube.

In the final analysis, the location of the deflection yoke on the tube throat will be dictated by conditions found in each individual case. The tube manufacturer identifies the theoretically correct location of the field, and also the length of the field, because the distance between the center of deflection and the start of the cone limits the maximum dimension of that half of the yoke. Information of this type is intended primarily for the equipment designer, but it is pertinent to us in connection with the matter of deflection-yoke replacement.

Another requirement is that the tube axis lie along the center of the magnetic field, so that the electron beam will be deflected symmetrically. Finally it is assumed that the two coils A-A' are identical and that equal numbers of lines of flux issue from each of the windings. Under such conditions, the field within the space may be assumed to be uniform, provided that the construction of the coils is such that the organization of the windings is the same in both.

Referring once more to Fig. 6-32, the axis of the tube or the undeflected path of the electron is X-X', X



Courtesy North American Phillips Co.

Fig. 6-33.—Outline drawing of Type 10BP4 electromagnetic deflection cathode-ray tube; note the position of the effective center of deflection.

being the point of origin of the beam entering the deflection area, and X' being the geometric center of the screen S. The two coils are L_d long, thus making the field of equal length, fringe fields being neglected. Finally, the center of the field is D_s distant from the screen.

When current flows through the coils in one direction the beam is deflected upward from its normal path X-X' and follows a new path, which we shall assume to be as shown, striking the screen at the point Z. This represents a displacement of the spot from X' to Z, corresponding to a deflection Y along the screen. The new path of the electron beam forms the deflection angle θ with the axis of the tube, and the beam is bent through a portion of a circle with a radius of curvature R. If the direction of the current through the coils is reversed, the deflection will be downward, as shown by the dashed lines in Fig. 6-32. The point where the beam strikes the screen at the limit of its downward excursion is Z', causing the screen deflection Y' and the deflection angle θ'. Except for the use of magnetic deflection, this illustration and its description resemble very closely the electrostatic system shown in Fig. 6-6A.

Magnitude of Deflection

Let us now examine the factors which determine magnitude of deflection. We can do this from various viewpoints. For the moment, let us exclude the length D_s and the screen deflection and examine the deflection angle θ and the radius R. Since we are working with an electron beam, its velocity is a very important item, but as will be seen later, changes in velocity do not influence the ultimate result to the same extent as in the electrostatic case. Nevertheless, the accelerating voltage does have an effect on sensitivity, in the same direction as in the electrostatic tube. The higher the beam velocity, the less the deflection per unit deflection force. So, for the moment, let us say that the accelerating voltage is fixed at some level, the exact value of which is unimportant.

Remembering the principles of magnetic deflection as outlined in Chapter 3, the deflection angle and the radius are functions of the magnetic field strength and the length of the field. In turn, the former is a function of the current through the coil and the number of turns in the coil. The current, in turn, is a function of the frequency and the magnitude of the deflection voltage applied to the coil, assuming fixed constants for the coil. As a matter of fact, we must make certain assumptions as we go along in this discussion.

First of all, we neglect the factor of frequency, although we recognize its effects, to eliminate one of the variables. Therefore, the current through the coil will be imagined as being d.c. Second, since the coil dimensions are fixed for any one system, the number of turns is fixed and so is the length of the coil. This leaves us with the current in the winding as the chief operating parameter. Since the intensity of the field is proportional to the current which produces it, both the deflection angle and the radius through which the beam is bent are determined by the amount of current in the deflection coils.

The consideration of current is sufficient if we are speaking about a deflection coil of known dimensions and known number of turns, but if we are thinking in general terms without having any specific coil in mind, stating the amount of current only is insufficient information. Then it becomes necessary to speak in terms of ampere-turns. The reason for this was discussed in connection with magnetic focusing and hence need not be repeated here. From the practical viewpoint, that is, working with existing deflection coils, the greater the current through the winding, the smaller the radius of curvature of the beam and the greater the deflection angle θ.

The Deflection Angle

Since deflection in everyday language usually means displacement of the spot both above and below, and to the right and left, of the undeflected spot position, the deflection angle θ associated with deflection in one direction, amounts to only one half of the total angle of deflection possible. Identifying the other half of the angle by θ', the complete description of the deflection angle, therefore, is $\theta + \theta'$. This is commercial practice and the angle referred to, in tube specifications, considers the center of deflection as the vertex and identifies the angle bounded by the maximum allowable swing of the beam.

Reference to the deflection angle affords an *impression* of the image size capabilities, but does not necessarily give information concerning trace dimension for the maximum deflection angle, because it does not take into account the distance to the screen. It stands to reason that for *unit over-all length,* the tube rated at the greatest deflection angle will result in the largest-size trace because it will afford the greatest deflection on the screen. For any one deflection angle, it stands to reason that the greater the distance D_s to the screen, the greater will be the deflection Y on the screen.

Deflection Sensitivity

The deflection sensitivity of a magnetic-type cathode-ray tube may be expressed in terms of the *deflection* on the screen per unit field strength, as for example in millimeters or centimeters per gauss or ampere-turn, or in terms of the *deflection angle* per unit field strength. Obviously, the first method takes tube length into account, whereas the second does not, although any numerical reference to a deflection angle implies a specific tube with a fixed geometry. The only difference between the two methods is that one gives the screen deflection for a specific tube, while the other requires projection of the deflection angle onto the screen, but is independent of the particular tube used. Neither case describes the deflection system other than to state the required field strength. Reference to required field strength is not general practice; however, in the tube specifications, usually only the over-all deflection angle is given.

From the practical viewpoint, the deflection system is a completely assembled unit, for which replacements are available, which are duplicates of the original, and are intended to be used with a specific tube. This is so for television receivers, or for any other complete assembly of equipment which makes use of a magnetic deflection yoke. The problems of design of such yokes are many, but since this is not a design text, such details are omitted and the interested reader is referred to the bibliography contained herein.

Concerning the deflection on the screen, it is expressed mathematically as

$$\frac{0.3 \times L_d \times D_s}{\sqrt{E_a}} \text{ cm per gauss} \qquad (6\text{-}11)$$

where L_d and D_s are in accordance with their meaning in Fig. 6-32 and E_a is the accelerating voltage applied to anode 2. Equation (6-11) assumes that the number of turns and the current in the winding is such as to develop a field intensity of 1 gauss per unit area.

Concerning the variables associated with deflection, the accelerating voltage becomes a part of such specifications, but it is to be noted that it appears as a square-root term. This means that doubling the accelerating voltage in a magnetic-type tube will have a smaller effect on the reduction of sensitivity than in the case of the electrostatic tube. To be exact, the former will cause only a 30 per cent decrease in deflection sensitivity, whereas doubling the second-anode voltage in the electrostatic tube reduces the sensitivity by 50 per cent. This relationship does not make the second-anode voltage any less important; it simply means that changes in this voltage have a reduced effect on sensitivity.

Deflection Factor

The deflection factor of magnetic-type cathode-ray tubes is related to deflection sensitivity just as in the case of the electrostatic-type tube, except that in the magnetic case it defines the current required in a standard coil to produce the 26° deflection angle. Such a standard coil has a specific number of turns and a specific physical length, since both of these factors play a dominant role in determining the amount of deflection of a beam of a certain velocity.

The Length and Location of Deflection Fields

Specifying the ampere-turns rating is one method of describing a specific coil, but two or more coils of widely different configuration may have the same number of ampere-turns. Thus the physical dimension of the coil relative to the length of the field it produces is an important point, because it determines the location of the deflection system around the neck of the tube with respect to the deflection of the beam. For example,

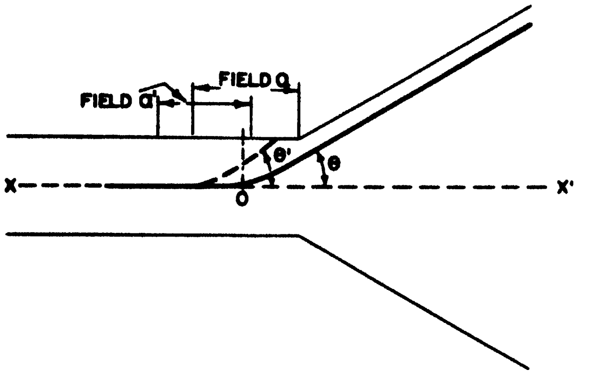


Fig. 6-34.—Here field *a* is properly positioned, while field *a'* is improperly positioned along the neck of the tube, since its geometric center does not coincide with the effective center of deflection, consequently the beam is deflected too early at an angle *θ'* and is blocked by the neck of the tube.

in Fig. 6-34 is shown that portion of the neck of the tube which joins the cone. Here field *a* is that due to a deflection coil which is properly positioned relative to the effective center of deflection of the tube. The length of the field has been arbitrarily chosen; in practice, it may vary from 1½ inches to perhaps around 3 inches as determined by the specific design. When located as shown, the field of length *a* is such that its midpoint coincides with the effective center of deflection of the tube and its endpoint terminates at the cone. The maximum deflection is indicated by *θ*.

Let us now imagine that the same deflection coil is moved slightly back toward the tube base, away from the cone, so that the field still has its original length, but now occupies the place shown as *a'*. If the intensity of the field is unchanged, then the angle $\theta' = \theta$, as before, and the deflected beam will be intercepted by the neck of the tube as shown by the dashed lines. Naturally, deflection angles less than the maximum shown will permit the beam to reach the screen, but the *location* of the deflection coil has introduced a definite restriction upon the maximum deflection angle and, therefore, on the magnitude of current which may be passed through the deflection coil.

For another example, assume that the deflection coil is one of different dimensions so that the field it produces extends over a length equal to 2*a*. Because of the increased length of this field, even though one boundary of it is located at the proper point along the neck of the tube, the center of the field is farther back along the neck than the effective center of deflection. Assuming a field intensity which will develop the same deflection angle as in the previous examples, *or even somewhat less,* the beam will not reach the screen; it will be blocked by the sidewall of the tube. Thus it is seen that both the location and the length of the deflection field have a bearing on the beam path, and hence on the utility of the tube in general.

These conditions should not be construed as indicating that the shorter field is preferable. Actually, the longer the field, the greater is the *deflection efficiency* because the beam is under the influence of the field for a longer period. It is also necessary to bear in mind that the beam being deflected is not a thin, round pencil of electrons, but has a relatively substantial cross section. Interference with a *portion* of the beam is possible, thus reducing the beam density at the screen and decreasing the light output.

Practical Deflection Yokes

The practical deflection assembly used with magnetic-type tubes consists of two pairs of deflection coils, one for the vertical deflection and one for the horizontal deflection. The assembly as a whole is identified as the *deflection yoke,* and is so designed that it fits fairly snug over the neck of the tube and is movable, so that it may be positioned properly. The entire yoke may be tilted slightly relative to the tube axis in order to compensate for what possible misalignment may exist within the gun assembly of the tube, thus enabling correct centering of the beam on the screen.

Seldom does a fixed-position deflection yoke mount a single deflection system, that is, a single pair of coils, since in a fixed position, it is suitable only for the display of a scalar quantity (one which has magnitude, but no direction). Used with another yoke which would mount the other pair of coils, it would mean lateral displacement of the deflection system along the path of the electron, as in the electrostatic case, with its corresponding disadvantages. However, as we said, such an arrangement is seldom used, because the conventional construction affords the necessary symmetry in the resultant field.

Relative to the physical design of the deflection coils, several varieties have been used. These may be divided into four fundamental types, consisting of: 1. air-core coils with lumped windings; 2. random-wound coils using a core of magnetic material; 3. multilayer coils using a core of magnetic material; 4. pi-wound coils using a core of magnetic material. Of these four varieties, the first has been quite popular in the postwar television receivers, although at the time of this writing, the second variety is receiving much attention. The other two types, as well as the first kind, were used in military equipment during World War II.

Each of these types possesses its salient features. The air-core winding affords operation over a wider frequency range, whereas the use of a magnetic core by concentrating the flux lines affords greater operating efficiency, and therefore, lowers the power requirements and the cost. The pi-wound coil in turn affords a high value of inductance with minimum distributed capacitance and tends to reduce the *recovery time,* by which is meant the time required for the coil to dissipate the energy supplied to it during a sweep cycle; also, it reduces the tendency toward self-oscillation.

Air-Core Deflection Coils

The air-core type of deflection coil used in television receivers and other kinds of equipment is shown in Fig. 6-35A. This illustration depicts one pair of such coils which form a system for deflecting the beam along one coördinate axis. A complete two-dimensional deflection system utilizes four such coils. Each is wound as a flat, rectangular-shaped coil with lumped turns. Then it is formed into a semicircular shape with the ends turned up. The effective legs are the two long sides of each coil, designated as *1* and *2* for coils *A* and *A'*. The radius of the circle formed between the coils is determined by the outside diameter of the tube neck over which the complete yoke is placed.

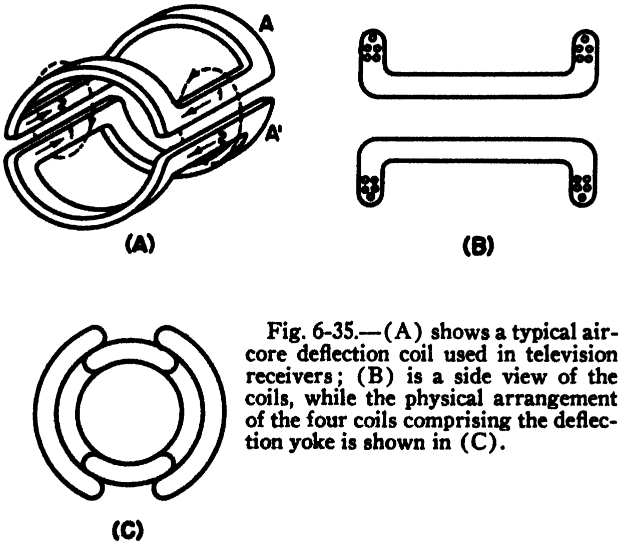


Fig. 6-35.—(A) shows a typical air-core deflection coil used in television receivers; (B) is a side view of the coils, while the physical arrangement of the four coils comprising the deflection yoke is shown in (C).

As can be seen, one leg of each coil is placed adjacent to the other, and by suitable connection, the current in these adjacent legs is made to flow in the same directions, as indicated by the arrows, thus causing the field of one to aid the other. A unidirectional field is, therefore, produced within the hollow core of the coil. This is indicated by the arrows on the dotted representations of the field. By turning up the short legs of the complete windings, the fields surrounding the long legs buck the fields surrounding the short legs; the limits of the field active on the electron beam are, therefore, well defined, and the fringe field is small. The side view of such a pair of coils is shown in Fig. 6-35B.

The physical arrangement of the four coils which comprise a complete deflection yoke are shown in Fig. 6-35C. As can be seen, one pair of coils is within the other, the active legs of the outer pair being located at the open areas of the inner pair. A photograph of a complete yoke is shown in Fig. 6-36. The wingnut on the yoke cover is used for tightening the assembly on its frame, thus keeping it steady in its proper position along the neck of the tube.

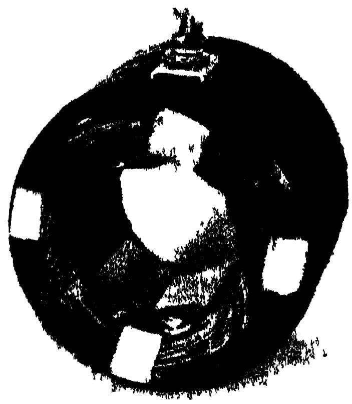


Fig. 6-36.—A photograph of the complete yoke shown in Fig. 6-35.

Iron-Core Deflection Yokes

Such units come equipped with a magnetic core made of a special kind of iron dust.[8] The appearance of the yoke remains substantially the same, although the performance is improved greatly, by attaining increased flux density where it is needed most. This lowers the value of current required by the deflection system. By the use of iron dust, eddy current effects are minimized, thus enabling beam deflection to start more rapidly. Finally, the greater efficiency in the deflection system permits the reduction of the required driving power from the tubes which feed the vertical and horizontal deflection coils.

Another type of deflection-coil arrangement employs a square-shaped core made of magnetic material, as shown in Fig. 6-37A. The deflection coils are mounted on the legs of the core, opposite legs carrying

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[8] Friend, A. W., "Television deflection circuits," *RCA Rev.*, March, 1947.

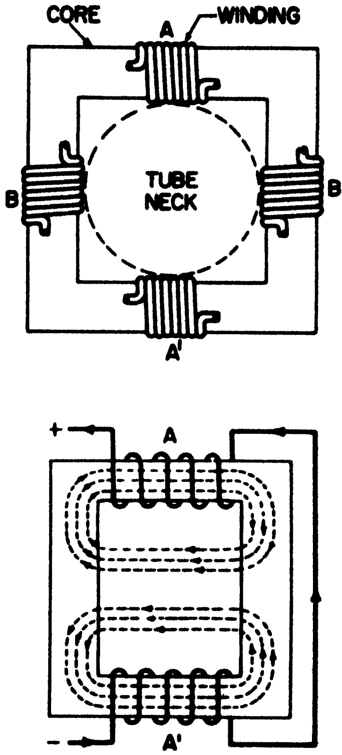

Fig. 6-37.—(A) shows a square-shaped core made of magnetic material; the connections and the field produced by a single pair of coils is shown in (B).

the two coils which produce deflection along the same co-ordinate axis. Each of these coils is made to occupy the entire leg, thus attaining symmetrical positioning of the coils and a uniform field within the hollow portion of the core. Inductive coupling is minimized, because equal emf's in opposite directions are induced in one pair of coils by the other pair of coils.

The flux lines associated with each pair of coils traverse for a portion of their path through the core, but complete their path by leaving the core and traversing the space between the coils. This is shown in Fig. 6-37B for a single pair of coils. It can be seen that the direction of the flux lines is opposite in the adjacent legs due to the series-bucking connection of the coils, and they leave the core at the midpoint. This illustration also shows how effectively zero inductive coupling between the pairs of coils may be accomplished by symmetrical placement of the coils. It is clear that the flux coil *A* links one half of the windings of the other pair *B-B'* (not shown in the figure), and thus develops an emf in coils *B-B'*. However, the flux from coil *A'* develops an emf in the opposite direction in the remaining halves of those two coils, as shown by the opposing arrows on the flux lines from *A* and *A'*. One offsets the other and the resultant emf is zero. In like fashion, coils *B-B'* react on coils *A-A'*, thus preventing mutual interaction.

Several ways of accomplishing like direction of the flux lines in the hollow of the core exist. It may be a series-aiding connection of similarly wound coils, a series-opposing connection of oppositely wound coils, or a parallel connection of two similarly wound coils. The field established by coils *B-B'* is similar to that of coils *A-A'*, except that the flux lines of *B-B'* are at right angles to those of *A-A'*, since the coils are positioned mutually perpendicular.

The efficiency of deflection coils of the type shown in Fig. 6-37 is reduced, however, because of two reasons. First, considerable space in the corners of the core window is wasted, as is that space occupied by the winding between the core and the tube neck. Second, there is an external flux developed outside of the core window which is wasted. Although the yoke may be shielded to lower this external flux, other disadvantages result from the use of the shield, so that this solution is very often impractical.

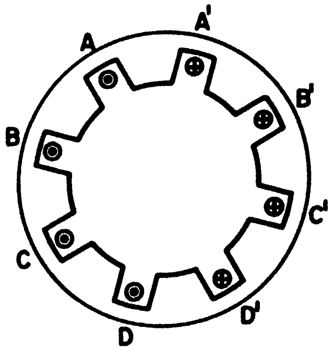

Fig. 6-38.—Cross section view of a deflection core made of slotted cylindrical laminations to improve the magnetic efficiency.

In order to improve the efficiency of iron-core deflection coils, yokes have been developed which utilize slotted cylindrical laminated iron cores.[9] Formed windings, similar in shape to those in Fig. 6-35A, are distributed in slots, much in the manner of motor windings. A simplified single-winding coil of this type is shown in cross section in Fig. 6-38. High orders of efficiency and sensitivity are claimed for this deflection-coil configuration, which is applicable to both high- and low-speed sweeps.

Push-Pull Deflection Systems

The types of windings described so far were intended for what is known as "single-ended feed systems." Coils may be wound for push-pull operation in either or both directions by simply placing two windings on each leg of the core. These windings are located

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[9] Woroncow, A., "Iron-cored deflecting coils for cathode-ray tubes," *Jour. IEE* (London), pp. 1564-1574, vol. 93, part IIIA, March-May, 1946.

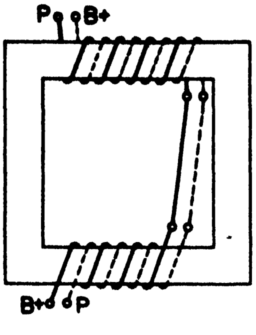

Fig. 6-39.—Deflection coil wound for push-pull operation; note that the inside winding on one leg is connected to the outside winding of the other leg.

on top of each other and are connected as shown in Fig. 6-39, where the inside winding on one leg is connected to the outside winding of the other. The connection between windings is such as to produce flux lines in the same direction in the core. The two terminals marked $B+$ are joined, since they are both connected to the source of plate potential of the driver tubes. The two windings marked P are connected to the plates of the push-pull driver tubes. With this arrangement, the "no-signal" plate current fields are balanced out and only the differential-signal plate current is responsible for deflection. The use of push-pull deflection is not limited to iron-core coils; it is just as readily applicable to air-core coils, and the arrangement of the windings is the same. Although only one set of deflection coils is illustrated, push-pull windings can be used for vertical and horizontal deflection with equal facility.

Radial Time Base or Polar Co-ordinate Display with Magnetic Tubes

Up to this point, the trace produced on the screen by magnetic deflection of the beam was considered only along rectangular co-ordinates. Like the electrostatic system, magnetic deflection is also capable of producing a display organized along polar co-ordinates. Two methods of attaining this, which have been used in many cases, will be described.

One of these consists of the mechanical rotation of the deflection yoke, which is the equivalent of saying that the magnetic field hitherto assumed to be stationary relative to the tube axis, is rotated around the tube axis. This is illustrated in simple form in Fig. 6-40. Coils A-A' represent the elements of a single deflection field. Whether they are air-core or iron-core coils is of no consequence here. They are mounted on a yoke, which also bears the necessary gear arrangement and two slip rings for connecting the current source to the coils. A rotating device spins the yoke around the tube throat.

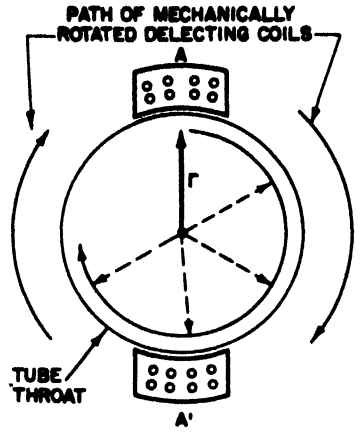

Fig. 6-40.—A rotating magnetic field produced by mechanical rotation of the deflection yoke.

If the polarity of a direct current is such as to cause deflection of the beam in the upward direction as shown by the line marked r, rotation of the yoke with the current in the coil held at a constant value will cause the beam to describe a circle as indicated by the dotted-line positions of the rotating deflection vector r. Obviously, the magnitude of deflection is the rotating radius of the circle described by the beam. Changing the value of current in the coil will change the magnitude of r, and therefore, the dimensions of the circle. If the coil current undergoes a continuous change in value from zero to a maximum and this is repeated periodically, the path followed by the beam will be a spiral, and such a trace will appear on the screen.

By synchronizing the rotation of the deflection yoke with some signal source which will control the intensity of the beam, as for example a rotating antenna, the movement of the spot along the screen surface will maintain angular agreement with the rotating antenna, and signals picked up by the antenna can, after proper amplification, present azimuthal information concerning the signal. If the amplitude of the rotating radius bears a relation to distance traveled by the search signal radiated by the rotating antenna, the display on the screen of the tube will indicate distance to the unseen target which has returned the echo of the search signal to the antenna. This is an elementary explanation of the Plan Position Indicator or PPI tube used to indicate the distance to the target, as well as its direction relative to the point of interception.

A radial time base or polar co-ordinate form of display can be developed without mechanically rotating the deflection fields. By utilizing the conventional deflection yoke, fixed in position and containing coils for vertical and horizontal deflection, and supplying these coils with currents which are 90° out of phase, the resultant deflection will impart a rotary motion to the beam, similar in behavior to that developed when two voltages 90° out of phase are applied to the deflection systems in an electrostatic-type tube.

Effects of Nonuniform Magnetic Deflection Fields

A number of conditions associated with the lack of uniformity of magnetic deflection fields have a great bearing on the successful use of this type of deflection. These are apart from the constants of the deflection systems, which are the only determining factors for the ideal case of uniform fields. In practice, one may experience nonuniform fields due to a wide variety of conditions, and it is these and their effects which interest us now.

These effects may be classified under two general headings: 1. defocusing, and 2. pattern distortion. Both stem from the same conditions, although the results may manifest themselves in different ways.

Deflection Defocusing

Deflection defocusing does not necessarily require the presence of two crossed fields. The action of a single field, either the vertical or the horizontal, may be such as to cause defocusing of the beam. The action arises from two related causes. One is the fact that the beam itself is not of infinitely thin cross section, but has a substantial area. If this beam is acted upon by a uniform field as shown in Fig. 6-41A, all parts of the beam are deflected by the same amount and the shape of the beam is not changed during the deflection process.

But if the field is not uniform as shown in Figs. 6-41B and C, all parts of the beam will not be acted on in similar manner and the shape of the beam will be changed during the deflection process. It may be changed into an ellipse which has its long dimension in the direction of the deflection, as would be the case in Fig. 6-41B, or it may have somewhat the same shape with the short dimension being in the direction of the deflection, which would be the case for the field shown in Fig. 6-41C.

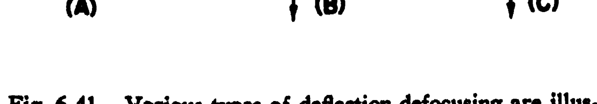

Fig. 6-41.—Various types of deflection defocusing are illustrated. (A) shows a uniform field acting upon the beam with no distortion resulting; defocusing produced by a nonuniform "pincushion" field is shown in (B), while the distortion resulting from a nonuniform "barrel" field is shown in (C).

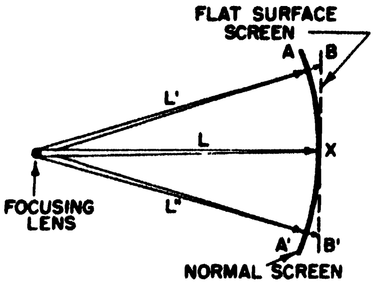

Fig. 6-42. — Spot-defocusing produced by a flat-face tube screen.

Correction of such defocusing is a matter of eliminating the conditions which give rise to the nonuniformity of the field, which will be described later in this chapter.

Flat-Face Tube Screen Defocusing

While on the subject of defocusing, it might be well to mention the effect of a flat-face tube screen as compared to one which has the proper curvature, although this is not related to deflection defocusing. With the image focal length of a focusing system adjusted to bring the beam to a focus at the center *X* of the screen *A-A'*, as shown in Fig. 6-42, the beam will remain in focus because the curvature of the screen conforms with the arc completed by the deflected beam. Stated differently, the radius of the screen curvature corresponds to the distance *L* from the center of deflection to all useful parts of the screen surface. Therefore, the focusing adjustment makes the screen the point where the second beam cross-over point is located. This is shown more clearly in Fig. 6-43, where two points have been selected for examination. One of these is the

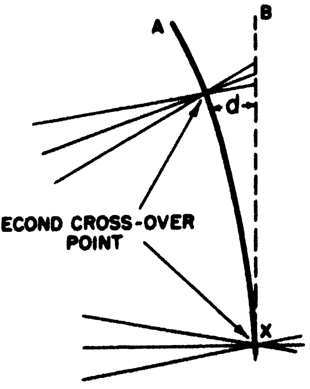

Fig. 6-43.—Defocusing produced by a screen of improper curvature; the second cross-over point of the beam occurs correctly at point *X* of the center of the screen; at the limits of deflection, however, the second cross-over point occurs too early at *A*, resulting to a diverging spot, as the beam travels through the distance *d* to the screen *B*.

center *X* on the screen surface, and the other is one limit of the beam swing. *A* and *B* are two possible screen surfaces. Point *X* is the center of both screens; therefore, the second beam cross-over point would occur on both screens at the same place and the image would be properly formed. However, the point on the screens corresponding to maximum deflection of the beam in the upward direction will coincide with the second cross-over of the beam on *Y* for screen *A,* thereby retaining the condition of focus, but will be ahead of the cross-over point for screen *B,* where the beam was again beginning to spread.

Since the limits of screen *B* are located in front of the second beam cross-over point, the image on the screen will not be in focus. Finally, the distance of *d* between the location of the second cross-over or image point at screen *A* and screen *B* becomes progressively greater as the beam advances from the center toward the limits of screen *B,* leading to an increasing amount of defocusing. This will be true for any direction of deflection away from the center of the screen.

Deflection Distortion

Two prominent types of pattern distortion due to nonuniform fields and related to deflection defocusing are known as *pin-cushion* and *barrel* distortion. They represent the type of pattern created by correspondingly shaped nonuniform fields in contrast to a nominally square pattern developed when the two crossing fields are uniform. They are shown in Figs. 6-44A and B. Normally the lines of flux of each field are at right angles to each other throughout and the resultant deflection is proportional to the amplitude of the vertical and horizontal components.

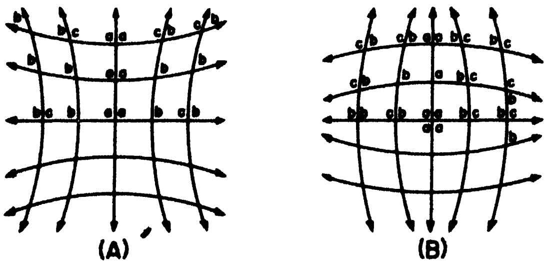

Fig. 6-44.—Distortion due to nonuniform fields; (A) shows a "pincushion" type field, and (B) a "barrel" type field.

But in fields of this type where the angle formed by the component forces may be acute or oblique, the resultant deflection force is no longer proportional to the individual component amplitudes only, but involves the angles formed by the intersecting fields. For example, two forces of equal magnitude acting at right angles to each other will develop a resultant of a certain magnitude as shown in Fig. 6-45A. However, the same two forces acting at some angle less than 90° (acute angle) will develop a greater resultant force than if the angle were 90°, whereas two like forces acting at an angle greater than 90° (oblique angle) will develop a resultant force less than when the angle is 90°. These are shown in Figs. 6-45B and C. In both instances, the resultant is obtained by the parallelogram method of vector addition.

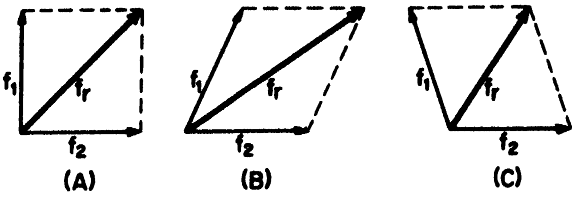

Fig. 6-45.—Change in the magnitude of the resultant deflection due either to acute angles between the field lines (B) or oblique angles (C). These distortions are produced by nonuniform "pincushion" and "barrel" type fields.

If you will examine the two fields of Fig. 6-44, the presence of all three types of angles between the flux lines will be noted. A few typical ones are indicated by the letters *a, b,* and *c,* where the meaning of each is in accordance with the illustrations similarly identified in Fig. 6-45.

The upshot of all this is that a normally square pattern may be changed to the types shown in Fig. 6-46A and B, the former being the pin-cushion variety and the latter being of the barrel type. The forms of field distortion shown in Fig. 6-44 are only two of very many possible types, since the lines of flux due to each field need not possess the symmetry shown. Varying degrees of tilt may exist between the axes, in which case, the pattern as a whole may be tilted, as shown in Fig. 6-46C, or possibly a curvature may be given to otherwise straight-line traces, etc.

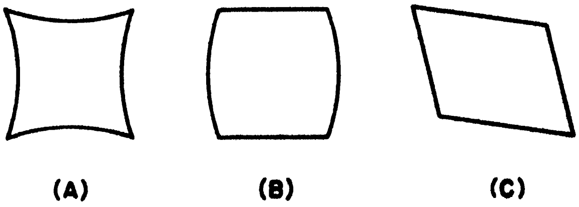

Fig. 6-46.—Common distortions resulting from nonuniform deflection fields; "pincushion" and "barrel" type distortions are shown in (A) and (B), respectively, while a tilting of the whole pattern is shown in (C).

Reasons for Nonuniform Fields

The possible reasons for nonuniform deflection fields are numerous. Anything which will tend to impair the axial relationship between the two deflection systems, so that they are not at right angles to each other will produce some form of distortion of the final trace on the screen. This may not be as severe as illustrated in Fig. 6-46, but it is generally noticeable. A tilting of the trace as a whole is a commonplace result.

The organization of the turns which comprise the windings, even in lumped coils, has a great bearing on the uniformity of the field. While the series connection between the two coils of a system will result in equal values of current through each winding, it does not necessarily assure that the resultant field will be uniform. The field due to each coil will be different, if the arrangement of the turns on each coil varies very greatly. This may give rise to curved traces or lines. A certain symmetry is maintained even under such conditions, but the required perpendicularity is not retained.

The presence of undesirable magnetic fields, due to other current-carrying devices, may interfere with the existing field, in such a way as to tilt the entire pattern, or cause a curvature of the trace. The lower the frequency of these stray fields, the more pronounced is the effect. Proper shielding of the interfering source, or reorientation of the source relative to the axis of the beam and the fields of the deflection system may correct the trouble. There are, of course, cases when the field is distorted deliberately in order to attain a special type of deflection, but these are so noted in the instructions which accompany such apparatus.

While deflection systems may not be identical, that is, the horizontal system may contain coils which differ from the design of the vertical system, the symmetry of the coils in each pair must be maintained. Naturally certain tolerances exist, but it is important that the rated tolerances are not exceeded.

Improper positioning, different lengths, or different number of turns on the coils of each pair on the square iron-core deflection yoke will also produce a nonuniform field. This may result in various types of curvature being given to the trace.

Rotation of the Pattern

In the event of interaction between the deflection and focusing systems whereby the deflection action takes place before the principal focusing action has been completed, the pattern as a whole will be rotated through an angular distance dependent on the amount of interaction. Since the action does not involve distortion, its correction is simply a matter of rotating the deflection system by the amount necessary to correct the radial component of velocity imparted to the deflected beam by the portion of the focusing field which overlaps it. Of course, the proper correction is removal of the overlap by separation of the two fields or by shielding one from the other. This may not be easy to do, in which case, rotation of the deflection yoke is the next best remedy.

Constants of Deflection Coils

Each winding of a deflection system is rated at a certain value of inductance. This information usually appears in service data accompanying the apparatus, especially television receivers. The usual frequency for checking the inductance is 1,000 cps. Since the values are not standard, little is gained by quoting actual figures. Moreover, the recommended method of checking television receiver deflection-yoke windings consists of a d-c resistance measurement.

In the case of military equipment, the inductance is given and the testing frequency is as stated above. Inductances vary over quite a range from about 50 millihenrys to as high as approximately 550 millihenrys.

As far as the inductance and the d-c resistance of the horizontal and vertical deflection coils used in television are concerned, the horizontal system, operating at a higher frequency than the vertical, employs coils with fewer number of turns. This reduces the inductance as well as the d-c resistance for a certain required impedance and corresponding deflection. The converse is true in the case of the vertical deflection coils. The operating frequency is lower, therefore, the inductance required for a comparable deflection is higher, and since this is attained by the use of more turns of thinner wire, the d-c resistance is greater.

Concerning the d-c resistance rating, a tolerance is to be expected. Whether or not a few shorted turns are determinable depends upon the over-all resistance and the accuracy of the equipment used. A discrepancy in d-c resistance may be overlooked, if the performance of the coils is satisfactory.

Also associated with each winding is the capacitance of the coil. The effective capacitance of the winding plays an important part in the recovery time of the coil, as described in the paragraph which follows, but this constant is not given in commercial equipment. It is, however, specified for military equipment, wherein the range of values extends from about 150 to 300 $\mu\mu f$.

Recovery Time

When current is made to flow through an inductor, energy is being supplied to that element and it appears in the field which surrounds the coil. Whether or not the energy, fed to the coil, can be dissipated in time when the current changes abruptly is a function of the recovery time of the coil, which depends on the design of the coil. This is a problem in deflection systems which utilize sawtooth-shaped currents, or in general, all currents which have very steep slopes. The recovery time of a coil is the time required for the field, due to a current, to collapse completely when the current ceases. The higher the repetition frequency of the cyclic change in current, the more difficult it is for a coil to recover in time.

To make the recovery time as short as possible, a damping resistance is connected across the coil. Such resistors will be found connected across each coil in the vertical deflection system. In the horizontal deflection system, the resistors are replaced by a more complicated tube system (damping tube), because of the problems imposed by the relatively high frequency of the current fed to those coils.

The value of resistance required across a coil to produce critical damping may be determined from

$$R = \frac{1}{2}\sqrt{\frac{L}{C}} \tag{6-12}$$

where

R is in ohms

L is in henrys

C is in farads.

Thus for a hypothetical coil rated at 100 millihenrys and an effective capacitance of 200 $\mu\mu f$, the equation would read

$$\begin{aligned} R &= \frac{1}{2}\sqrt{\frac{0.1}{200 \times 10^{-12}}} \\ &= \frac{1}{2}\sqrt{5 \times 10^{8}} \\ &= 11{,}200 \text{ ohms (approx.)}. \end{aligned}$$

Magnetic Shielding of Cathode-Ray Tubes

The shielding of cathode-ray tubes, especially the electrostatic variety, against external magnetic fields is a very important and interesting detail. It is a problem which faces every user and designer of equipment. Unfortunately, the penetration of magnetic fields, particularly those of low frequency can take place, even when the source of the interference is remotely located from the cathode-ray tube. That is why magnetic shields are found around the tube inside of the housing.

Inasmuch as it is impossible to completely shield a tube, that is, completely enclose it, avenues of field penetration are the front of the tube and the base end. In the main, it is the electron gun which requires the most effective shielding, but this does not mean that the portion of the tube between the gun and the screen may be ignored. Many types of cathode-ray equipment employ partial shielding, with the space between the gun and the screen exposed. This has been found satisfactory when the tube is contained within a housing of magnetic material which tends to act as a magnetic shield, but even such systems can be affected by magnetic fields originating from near-by sources. A soldering iron or a small current-carrying transformer being operated from a 60-cps source, and located in close proximity to the oscilloscope, will invariably produce a ripple on the trace.

Of special interest is the fact that the lower the frequency of the field, the more difficult is the shielding problem. Arrangements suitable for operation at 60 cps and higher fall short in effectiveness at frequencies of several cycles or less. Oscilloscopes intended for d-c and low-frequency a-c operation demand much more critical shielding than those intended for use at the higher frequencies.

The construction of the shield is very important regardless of its permeability rating. Holes and slits in the shield should be kept to a minimum. The orientation of a slit has a bearing on the magnitude of leakage from an interfering field; if the long side of the opening is in the same direction as the line of force, the shielding effectiveness is materially reduced. It is always advisable to close off all such openings, if possible. Joints in the shield are detrimental to its effectiveness. If possible, it should be made of a single piece.

A variety of magnetic shielding materials are available from different sources. Sometimes materials of lower permeability are adequate for the purpose, but it is preferable to use materials of the highest permeability, if the cost is not excessive.

The presence of interfering magnetic fields may be manifested in different ways. If the earth's magnetic field is interfering, it causes a steady deflection of the beam, which can be corrected by suitable adjustment of the centering controls, although proper shielding is the better solution. The earth's field may cause a tilt of varying degree dependent upon the orientation of the beam relative to the field, and the intensity of the field. The latter is a function of geographical location.

Magnetic fields from current-carrying conductors will cause a steady deflection if the current is d.c., and a ripple on the trace, if the current is a.c. A very significant detail relative to the latter is the frequency of the trace being affected by the interfering field. The higher the frequency of the trace, the less noticeable is the effect of the interfering a-c field, whereas the lower the frequency of the trace, the more easily is interference observed. Experiments with a number of different oscilloscopes have shown that the cross-over point for such effects is about 5 cps. When the trace moves at speeds corresponding to this frequency or lower, comparatively low intensities of low-frequency magnetic fields will alter the character of the trace.

Sometimes it is possible by orientation of the source of the interfering field to minimize the effect, although the best remedy is better shielding of the tube or shielding of the source. When the direction of the interfering field is such as to cause deflection in the same direction as the trace advances, the presence of the higher-frequency field is indicated by bands of varying intensity riding on the trace. If the direction of the field is at some angle with the direction of the trace, the ripples appearing on the trace will be tilted or will be at right angles.

The higher the frequency of the interfering field, the less noticeable the effect, other than for an apparent broadening of the trace due to the modulation by the external field. The frequency of the interfering field may be determined by a critical examination of the modulated trace; the number of ripples modulating a single cycle of the trace multiplied by the frequency of the trace is the frequency of the interfering field.

Concerning the materials suitable for shielding, two conditions must be taken into account. The essential requirement is high permeability, so that the path of least magnetic resistance is offered to the line of flux, thereby keeping the field from within the space surrounded by the shield; however, solid shields are best suited for d-c fields. Increasing the thickness of a solid shield does not necessarily increase its effectiveness in all cases.[10, 11] Shielding against transients or alternating fields is best accomplished by laminated, rather than solid, shields. Eddy-current ecects in solid shields of substantial thickness tend to prevent changes in magnetization of the material in conformity with the field.

The materials which are available on the open market under various trade names are alloys of nickel and iron of various percentages, sometimes containing, in addition, other ingredients such as a small amount of copper. Well-known trade names of such substances are "Mumetal," "4750 Alloy," "Highmu 80," the first being available from Allegheny Ludlum Steel Corp., and the latter two from Carpenter Steel Co. There are also other brands which have been extensively used.

Concerning laminated shields, a layer of copper between the layers of the magnetic material is very effective. Such a combination will be more effective over a range of frequencies than a single thickness of the magnetic material, and will, at the same time, shield against d-c fields.

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[10] Ellwood, W. B., "Magnetic shields," *Bell Lab. Rec.*, vol. 17, pp. 93-96, November, 1938.

[11] Gustafson, W. G., "Magnetic shields for transformers," *Bell Lab. Rec.*, vol. 17, pp. 97-99, November, 1938.

CHAPTER 7

SCREENS

The usefulness of the cathode-ray tube depends upon its ability to present a visual pattern upon the face of the tube, the part called the *screen*. Hence, the over-all performance of the cathode-ray tube not only depends upon the electron gun assembly, the deflection system, and the applied voltages, but also, in the final analysis, upon the screen.

The action of the screen is not based solely upon electrical and magnetic principles as are the other elements of the cathode-ray tube, but involves the process of changing electrical energy into light energy. The medium for this transfer of energy is a chemical substance which functions as the screen itself. This screen material, also commonly known as a phosphor, is applied to the inside face of the cathode-ray tube. How well the screen converts the electrical energy into light energy depends primarily on the type of material used for the screen and also on the velocity of the electron beam. Although there are a number of different types of screens used today, in reality there are not too many chemicals that will emit light when bombarded by a stream of electrons.

In brief, the basic action of the electron beam on the screen is as follows: the electron beam hits the screen with a high velocity. Due to the chemical nature of the phosphor and the energy of the beam, the screen will emit light upon bombardment by these electrons. The color of the light will depend upon the type of phosphor used; the screens used today emit such colors as blue, green, and yellow, as well as others. After an initial build-up time, the screen will continue to glow at the same color and the same intensity, so long as the electron beam is hitting the screen. After the removal of the electron beam, the screen will continue to glow for some finite period of time. The persistence of this *afterglow* varies for different types of screens. The degree of persistence is usually divided into three classifications: short, medium, and long. The color of the light emitted during the afterglow may be the same as that during actual bombardment by the electrons, or it may be a different color. Let us first analyze what happens when the electron beam hits the screen of the cathode-ray tube.

Electron Action

We already know about the various factors which influence the speed of the electrons traveling to the screen, their deflection, and their focusing. After the electron beam leaves the deflection field, the only other factor that may influence it is the post-deflection accelerating anode, also known as the intensifier electrode or third anode. This electrode, as you will recall, only functions to accelerate the speed of the electron flow in reaching the screen. When the electron beam hits the screen, two fundamental actions occur. One involves the return path of the electrons, and the other refers to the manner in which light output is produced by electron bombardment. We will consider the former action first.

The Electron Path

A number of theories about the return path of the electrons have been current in the past. It is generally believed now that several processes are involved. Such factors as the secondary emission of the phosphor and glass base, the degree of conductivity of the phosphor, the velocity of the electron beam, and hence the second-anode or intensifier-electrode voltage, all play an important role. The simplest way, however, of summarizing these and tracing the return path, after the electrons hit the screen, is by means of the concept of *secondary emission*. By secondary emission we mean the release or emission of electrons from the screen material due to the impact of other electrons.

All substances consist of atoms which contain positively charged particles (protons), neutrons having no charge and about the same mass as protons, and negative charges (electrons). A certain amount of energy is required to break an electron away from the atom due to the electrostatic attraction between the negative charges and the positive nucleus. Electrons traveling at high velocities can strike a surface with such force (the traveling electron has a great deal of kinetic energy) that one or more of the electrons bound to an atom in the material may be forced out, a process which is called "ionization." The initial electrons striking the surface are usually called *primary electrons* and those released from the material are called *secondary electrons*. The amount of secondary electrons released is not necessarily equal to the number of primary electrons hitting the screen. The ratio between these two is primarily dependent upon the kinetic energy of the electron beam and the type of screen material used. When the number of secondary electrons released is equal to the number of primary electrons, the screen surface is said to be at an *equilibrium potential.* (The amount of charge leaving equals the amount arriving for a fixed time interval.)

Thus we find the screen acting as a virtual cathode due to the secondary emission. Let us find out where these secondary electrons go. The majority of the secondary electrons are attracted to and divide up between the near-by positive electrodes, which would be the second anode and the intensifier electrode, if one is employed. Some of the secondary electrons, however, are also attracted back to the screen itself in an effort to maintain the screen at a voltage equilibrium. This latter action is best explained as follows: since we are dealing with a chemical material which is not pure and normally does not possess a smooth surface, various parts of the screen will act differently with respect to the number of secondary electrons emitted. This means that as the primary electrons strike different parts of the screen, these parts may exhibit differences of potential between them, with one being slightly positive with respect to another. This positive portion will absorb enough of the secondary electrons, which are moving at a slow velocity, to prevent any two near-by parts of the screen from differing in potential to any appreciable extent.

Producing the Light Output

The process of producing light output by electron bombardment of the screen material is quite complex. It is beyond the scope of this chapter to describe the complete action, as this would involve intricate details of atomic physics. The following analysis is a *simplified* version of what basically occurs, whenever light is produced due to electron bombardment of atoms.

In order to understand the basic phenomena, we must first know something about the fundamental structure of an atom. An atom consists of positive and negative charges, and neutrons. The positively charged particles and neutrons are all grouped together and are referred to as the *nucleus* of the atom, and the negative charges electrostatically attracted to the nucleus are called *electrons.* In a neutral atom, the number of positive and negative charges are the same. There are a number of fundamental ways in which the atomic structure of the elements may be pictured. In Fig. 7-1 we have illustrated one basic method. In any atom, no matter what the element, the nucleus with its positive charges occupies the center of the atom. The electrons are distributed about this nucleus in *shells.* These shells are not physical lines and hence are shown dashed in the drawing. The shells are concentric with the nucleus and the distance separating them decreases, the farther away they are from the nucleus. The shell nearest the nucleus is called the *K* shell, the next is called the *L* shell, the next the *M* shell, and so on. Thus, we find that there are seven active shells in which the electrons of all existing chemical elements may be oriented. In the drawing of Fig. 7-1, only the first five shells are illustrated.

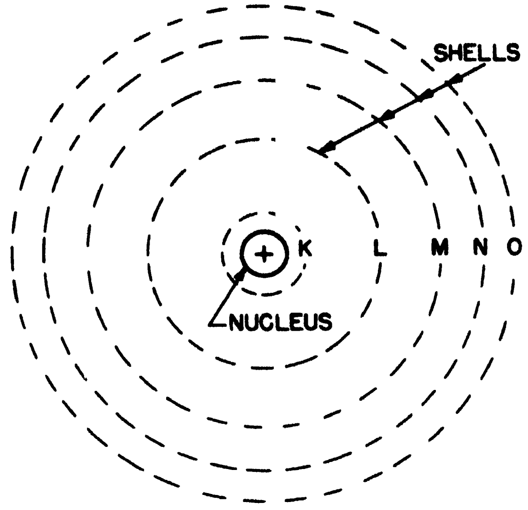

Fig. 7-1.—Basic structure of the atom, showing the positive nucleus in the center and unfilled electron shells.

The number of electrons in an atom depends upon the element referred to. For example, the element, lead, has 82 electrons, tin has 50 electrons, cobalt has 27 electrons, sodium has 11 electrons, and hydrogen has only one electron.[1] Not all the shells are occupied by electrons. Those nearest the nucleus are occupied first. No matter what the element is, the electrons distribute themselves in a specific manner in certain shells.

As an example, let us take the element sodium (chemical symbol Na) and see how the electrons are distributed. Refer to Fig. 7-2. Sodium has eleven electrons, which also means that the nucleus has a +11 charge. Two electrons are found in the *K* shell, eight electrons in the *L* shell, and hence only one electron appears in the *M* shell, as is indicated in the drawing. Note that the shells after the *M* shell are not occupied. When two electrons occupy the *K* shell of any atom and eight electrons the *L* shell, these shells are said to

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[1] The atomic number of an element tells us how many electrons the atom possesses.

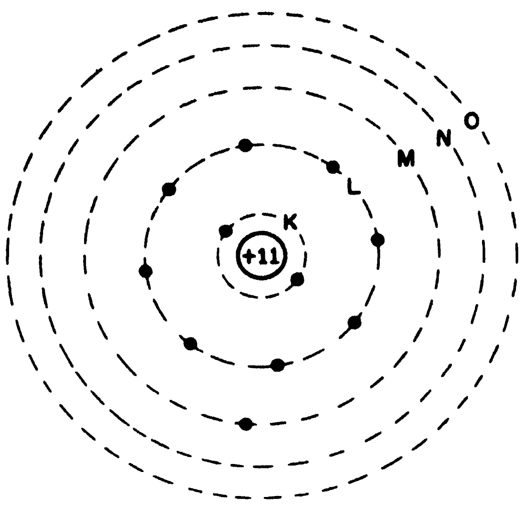

Fig. 7-2.—Structure of the neutral sodium atom, showing the nucleus which contains 11 positive charges and the 11 electrons distributed through the *K*, *L*, and *M* shells.

be filled. When a shell is filled, it is in a highly stable condition, that is, it is difficult to lose an electron from the shell and impossible to gain one. When a shell is not filled, such as the *M* shell in the sodium atom, it is relatively easy to gain or lose electrons from this shell.

With this fundamental understanding of the structure of an atom, we are now ready to study the basic manner in which light output is produced. When the electron beam hits the screen, we mentioned that one of the actions that occurs is secondary emission. This phenomenon actually causes removal of electrons from the atoms of the screen material. The other action, that causing production of the light output, primarily depends upon the *temporary* displacement of electrons from an unstable or unfilled shell.

Let us use the sodium atom once again for illustrative purposes and assume that as the electron beam strikes one of these atoms, the electron in the *M* shell (which is unstable) is forced out due to the impact of the beam. Furthermore, let us assume that the impact is not strong enough to completely remove this electron from the atom itself, but merely displaces it toward another shell. The electron is normally displaced toward an outer or unoccupied shell, where it possesses greater energy. In being displaced toward these outer shells, the electrons generally have a tendency to be temporarily (an extremely short time) seated in one of them.

The electron may be displaced toward the next outer shell or pass through one or two shells before temporarily resting in another shell. The exact shell the electron will go to depends upon the amount of energy it absorbed due to the impact of the electron beam. An electron in one of the outer shells has a progressively greater potential energy, the greater its distance from the nucleus, since more work is initially required to place it in one of the outlying shells against the attraction of the positive nucleus; for this reason, the outer shells are often called "higher energy levels." The electron dislodged by impact will go to a shell, where its energy is just equal to the amount of work which has been done upon it by the bombarding electron.

Because of the attraction by the nucleus, the electron immediately upon reaching some outer shell or energy level is attracted back to its original shell of lower energy, which is its normal position. Since the electron previously absorbed energy, and energy can neither be created nor destroyed, the electron in falling back to its original shell must give up this excess energy. It does so in the form of radiant power.

The important fact in this release of radiant energy is that it is in the form of electromagnetic waves and hence possesses a specific wavelength. Light waves are electromagnetic waves and are part of the electromagnetic spectrum. They occupy, however, only a small portion of this spectrum. Materials used for screens of cathode-ray tubes are so chosen that they will radiate energy, due to electron bombardment, in the visible light part of the spectrum.

The exact frequency of the radiant energy emitted depends upon the difference in energy of the shells through which the electron passes. Going back to the sodium atom again, let us assume two cases—one where the electron, originally in the *M* shell, is displaced to the *N* shell and then returns, and the other where the electron is displaced two shells outward to the *O* shell and then returns. Let us analyze the first case. Refer to Fig. 7-3A, which is the section of the sodium atom of interest to us. When in the *M* shell, the electron possesses a certain amount of potential energy, call it W_M and when displaced to the *N* shell, it possesses a greater potential energy, call it W_N. The difference between these two potential energies is directly related to a frequency of the spectrum as indicated by

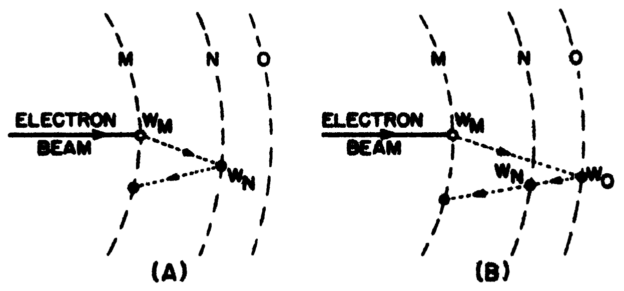

Fig. 7-3.—Electron bombardment of a sodium atom. (A) indicates electron knocked out of *M* shell to *N* shell; (B) shows the electron displaced to the *O* shell.

the following formula

$$W_N - W_M = hf_1 \qquad (7\text{-}1)$$

where f_1 is equal to the frequency radiated and h is a dimensional constant (called Planck's constant), which is equal to 6.624×10^{-27} erg-seconds. Solving equation (7-1) for f_1, we find

$$f_1 = \frac{W_N - W_M}{h}. \qquad (7\text{-}2)$$

The frequency in this case, therefore, is equal to the difference in energy between the adjacent energy levels divided by a constant. When the electron that has been displaced to the *N* shell returns to its original *M* shell, the radiant energy that it releases produces only one frequency line in the visible part of the spectrum.

Now let us take the second case, where the electron is displaced two shells outward to the *O* shell as indicated in Fig. 7-3B. In this case, when the electron is in the *O* shell, it possesses a greater potential energy, call it W_O, than when in either the *M* or *N* shell. When this electron travels back to its original shell, it releases a certain amount of radiant energy in passing from shell *O* to *N* and a different amount of radiant energy in passing from shell *N* back to its original shell *M*. In each case, the radiant energy released has a *different* frequency. In going back from shell *O* to shell *N*, the frequency f_2 is

$$f_2 = \frac{W_O - W_N}{h} \qquad (7\text{-}3)$$

and in going back from shell *N* to shell *M*, the frequency is the same as equation (7-2). The electron can also return directly from shell *O* to *M*, in which case still another frequency is radiated. Which transition will actually take place is determined by the so-called *selection rules* and depends on a number of complex factors, such as the type of circular or elliptical orbit the electron rotates in, the spin around its axis, etc. The chemicals chosen for the screen materials are such that although the frequencies radiated differ from one shell to another, they generally fall within the visible part of the spectrum. Consequently, no matter what shell the electron is displaced to, the phosphor used as the screen is such that the radiant energy released when the electron returns to its shell has a frequency *range* within the light part of the electromagnetic spectrum.

Electron Returning to an Ionized Atom

It is possible for electrons originally in an unfilled shell, when bombarded, to be completely knocked out of the atom, i.e., past all energy levels, so that there is no longer any attraction to the nucleus. When an atom loses an electron in this manner, it is said to be in an ***ionized*** state. If an atom loses two electrons from its unfilled shell, it is said to be doubly ionized.

When an atom is in an ionized state, it can receive any free electrons, not necessarily those it originally lost, to get back into a neutral state. Slow-moving secondary electrons that hover around the screen material may fall into an outer energy level of some ionized atom to make it neutral again. When an electron drops into the outermost energy level of an ionized atom, it is attracted by the nucleus toward the unfilled shell. In the process of falling back through the various levels, the electron radiates energy, most of which falls into the visible light region of the spectrum because of the nature of the screen material. Thus, we see that light output is also produced by electrons finding their way into the unfilled shells of ionized atoms.

The Electromagnetic Spectrum

We mentioned that light is an electromagnetic wave and hence occupies part of the electromagnetic spectrum. The electromagnetic spectrum is an orderly numerical arrangement of the frequencies of electromagnetic waves known thus far. This spectrum extends from frequencies as low as 10 cps to those as high as 10^{14} megacycles per second; theoretically, there is no upper limit, however. This means that the wavelengths of this spectrum extend from 30 million meters to 3×10^{-10} cm. Audio frequencies, radio waves, infrared waves (those which contain radiated heat), light waves (visible radiations), ultraviolet waves, X rays, gamma rays, and the cosmic radiations are all part of the electromagnetic spectrum.

That part of the electromagnetic spectrum in which we are interested is often called the ***visible spectrum*** and lies between the infrared and ultraviolet regions of the spectrum. It is an extremely thin strip lying approximately in the midregion of the electromagnetic spectrum. For radiations of very short wavelengths, there are two special units conventionally used for designating the wavelength. A wavelength of 10^{-4} cm is called the *micron* and designated by the Greek letter μ. Thus a wavelength of 0.03 cm is equal to 300 microns, often written as 300μ. The other unit used for still smaller wavelengths is called the *Angstrom unit* and corresponds to a wavelength of 10^{-8} cm. This unit is designated by the symbol Å. Thus, a wavelength of 0.03 cm is equal to 3 million Angstroms or 3×10^6 Å.

The visible spectrum extends roughly from a frequency of approximately 4.3×10^8 Mc to 7.5×10^8 Mc, which is a very small portion of the electromagnetic spectrum. In terms of micron units, the wavelength of the visible spectrum extends from about 0.70μ to 0.4μ, or in terms of angstrom units, it extends from 7,000 Å to 4,000 Å. This light spectrum gradually

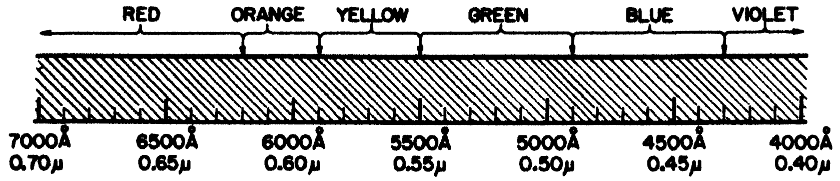

Fig. 7-4.—The visible light spectrum showing the approximate wavelength range of the chief color hues.

changes in color from a deep red at the low-frequency or high-wavelength end, as indicated in the spectral drawing of Fig. 7-4, passing through orange, yellow, green, and blue to violet at the short-wavelength or high-frequency end. There is actually no line of demarcation between any two colors as indicated on the drawing; each blends in with the two colors adjacent to it exactly like a rainbow. The extreme ends of the red and violet colors are not given any specific wavelengths, because there is actually no exact limitation as to the length of the visible spectrum.

When a displaced electron drops back to an unfilled shell of the atom, the wavelength of its radiant energy has to fall within the spectrum indicated in Fig. 7-4 in order to be visible.

Terminology

There are quite a few terms that help describe certain characteristics of the cathode-ray-tube screens. Definitions of these terms are necessary in order to follow the remainder of this chapter. This section will be devoted to these definitions.

Luminescent

Luminescent is the general term applied to any material that emits light when excited by some external source. The screens of cathode-ray tubes are luminescent materials because they emit light during bombardment by the electron beam.

Luminescence

Luminescence is the term usually applied to encompass all forms of radiation within the visible spectrum. There are many different types of luminescence. For example, bioluminescence refers to luminescence produced by living organisms (the firefly is an example of this) and chemiluminescence refers to luminescence produced by chemical reactions. Cathodoluminescence is the term applied to luminescence that is caused by the impact of electrons, produced by cathode rays, against some surface. This latter term is the one we deal with in cathode-ray tubes, where the surface is the screen of the tube.

Materials can luminesce *during* and *after* their period of excitation. There has been some discrepancy in the literature as to whether or not the term of luminescence shall also encompass light radiation after the source of excitation has been removed. Both interpretations have been used, but the most commonly accepted definition, and the one that we will use in this chapter, is that luminescence refers to visible radiation of the material *both during and after* the excitation period.

Fluorescence

The luminescence of a material *while it is being excited* is referred to as fluorescence. As far as cathode-ray tubes are concerned, fluorescence generally occurs during the time the electron beam is hitting the screen. The period of fluorescence of the screen, therefore, continues for approximately as long as the electron beam is active. Fluorescence can exhibit any color or combinations of color in the spectrum.

Phosphorescence

Phosphorescence refers to light emission which persists after the excitation has been removed. For cathode-ray tubes, this means that the screens will luminesce for a certain period of time after bombardment by the electron beam. The time limit of phosphorescence depends upon the type of material used. Phosphorescence can exhibit the same color as fluorescence, or it may be a different color.

The theory involving the production of light output due to phosphorescence of cathode-ray-tube screens is quite complex due to the impurities added to the screen material. It is basically similar to the process of light output during fluorescence, that is, radiant energy is emitted during the return of electrons to unfilled shells of the atoms. However, some of the electrons bombarded out of their unfilled shells by the electron beam do not immediately return to these shells. What happens is that these electrons pass into so-called *metastable* or *quasistable* states and remain there for a certain period of time. They generally return to the unfilled shells some time later by thermal collisions and then radiate energy in a manner similar to that occurring during fluorescence. Phosphorescence also occurs during fluorescence, but its glow is generally of a weaker intensity and not obvious, since it is blended in with the fluorescence. This is especially so if the fluorescence and phosphorescence are of the same color. Consequently, for all practical purposes, phosphorescence is said to exist after the period of fluorescence, that is, after electron bombardment has ceased.

Phosphor

The complete screen material made up of several constituents producing a characteristic luminescence is referred to as the *phosphor.* Occasionally it is interpreted to refer to the active light-producing part of the screen only. The term "phosphor" should not be confused with the chemical element *phosphorus.* Phosphors can be either fluorescent, phosphorescent, or both. Cathode-ray-tube screens are generally of the latter type.

Persistence

The term persistence refers to the duration of the period of phosphorescence or afterglow. As mentioned at the beginning of this chapter, the cathode-ray-tube screens are classified into three main groups—namely: *short-, medium-,* and *long*-persistence types. There is no fine line of demarcation between these groups. The period of phosphorescence for short-persistence screens may last anywhere from a few microseconds up to one millisecond, that for medium persistence screens generally falls between a few milliseconds and as high as one or two seconds, and any persistence longer than this amount falls into the long-persistence screen category. The above time limits only take into account the degree of phosphorescence noticeable by the human eye, and sometimes other definitions are encountered.

Luminous Intensity

In connection with cathode-ray-tube screens, it is important to know something about the light output. There are a number of technical terms relating to light with which we should be familiar. One of them is luminous intensity, which is defined as the amount of light flux radiated per unit solid angle, as seen from a point source of light.

Candle

The standard for luminous intensity is known as the candle. The source of light selected to define this standard is determined by specially calibrated systems throughout the laboratories of the world. Previously the international candle unit was the only standard, but the new one used in the United States is somewhat smaller than the international candle.

Candlepower

Candlepower is the luminous intensity of a light source expressed in candles. The abbreviation cp is used for candlepower.

Lumen

A light source emits light energy at some specific rate. The term *light flux* or *luminous flux* is used to specify the rate of light output from a source. A lumen is the unit of measurement of light flux. It is defined as the amount of light flux that a standard one-candlepower point source radiates through a unit solid angle. The greater the number of lumens measured on a surface, the greater is the light intensity at that surface. When we say a surface has greater illumination than another, we mean that the amount of light flux or lumens per unit area is greater.

Foot-candle

The foot-candle is a unit of illumination which is equal to one lumen per square foot. This is used only for the English system. The corresponding metric unit (lumens per square meter) is called the *lux.*

Foot-lambert

Brightness is the luminous intensity per unit area of a surface viewed from a specific direction; i.e., the illuminated area must be projected on the given direction. The *lambert* is the unit of brightness; it is the average brightness of a surface emitting or reflecting one lumen per square centimeter. It is also equal to $1/\pi$ candles per square centimeter. This definition is for the metric system. For the English system, the unit of average brightness is called the *foot-lambert.*

Writing Speed

Writing speed or writing rate refers to the linear velocity of the trace over the screen, usually stated in inches or centimeters per microsecond. The normal reference is with respect to sinusoidal input signals.

Current Density

The current density refers to the amount of beam current per unit area at the screen of the tube. It should be remembered that the beam current returns primarily through the anode and intensifier electrode (if the latter is used).

"Activator" Content of Screen Material

It is important to know something about the general chemical makeup of the screen material for a better understanding of the luminescence of the different screens. In the beginning of this chapter, we indicated how light output, due to electron bombardment, is produced by the giving up of radiant energy of a displaced electron in falling back to its previous position. The actual process is much more complicated than this simple analysis would indicate, because of the complex nature of the screen material. However complicated the process may be, it involves the production of radiant energy that falls for the greater part within the visible portion of the electromagnetic spectrum.

It is said that the chemical compound used as the phosphor when considered in pure form (with no other chemical impurities added) cannot luminesce because of the crystalline nature of the material. The ability of certain chemicals to luminesce is, in most cases, primarily dependent upon some impurity in the material, which is called the *activator*. The activator helps start the luminescent action within the base chemical and greatly increases its efficiency. This activator action in phosphors is analogous to a catalytic agent used to speed up some chemical reactions. On the other hand, it is also said that there are some phosphors which in their pure form *can* fluoresce without any added impurity or activator in the material. Although no direct chemical impurity exists in these, it is generally believed today that atoms of one of the metallic elements making up the chemical phosphor are situated in spaces between molecules of the material, and it is these atoms which are said to perform the function of activators.

Most of the activators used today are metallic elements, although there are some other types of elements and compounds that are used as activators. The most common type of metal activators used today are silver, copper, and manganese. Iron and nickel are also used as metallic activators, but not to the extent that the others are. *The type of activator used determines the fluorescent color of the phosphor.* This is definitely proven by one of the most common types of phosphors, namely, zinc sulfide. When this compound is activated with silver, it will fluoresce a blue color, when activated with copper, the fluorescent color will be blue-green, and when activated with manganese, the color will be orange. Pure zinc sulfide alone produces a pale blue color.

Besides the *type* of activator determining the fluorescent color, the *quantity* of the activator material also plays a very important role. If the amount of silver that gives a blue fluorescent color is increased slightly, the fluorescent color can be made to shift to red. This means that the wavelength of the radiant energy is increased. If too much of an activator is used, it is very possible that, instead of functioning as an activator, it will have just the opposite effect and prevent fluorescence. If used in this way, the ingredient is often referred to as a *quencher* to distinguish it from the term *activator*.

Only a very small amount of an activator is required to help produce fluorescence. Most fluorescent compounds as found in nature contain a certain amount of impurities. One or more of these impurities can easily act as an activator. Since only a very small quantity of the activator is usually required, it often happens that there is too much of some impurity in the natural phosphor and, instead of functioning as an activator, it performs as a quencher. Consequently, in the process of manufacturing screen materials out of natural phosphors, the phosphor is first put through a refining process to remove the impurities and then the required amount of activator is added to the purified phosphor. Phosphors are also synthetically manufactured, with the activators being inserted during the process of manufacture.

More than one activator can be used with certain phosphors. In cases such as this, two different colors can be generated during moments of fluorescence and may appear as a single blended color. It is during the periods of phosphorescence that these differences in color become obvious. Thus we see that the activators also have an effect upon the color of phosphorescence.

The usual fluorescent and phosphorescent colors used for cathode-ray-tube screens employed in oscilloscopes, television equipment, and radar apparatus are green, white, blue, orange, yellow, red, blue-green, and blue-white. At the time of writing, the only standardization relating to phosphors is that referring to tube type. The RMA standardization is that the third symbol of the cathode-ray-tube type shall be the letter *P* followed by a number, both of which designate the screen characteristic. Many such *P* numbers do exist, such as P1, P4, P7, etc., and the terminology is used by most manufacturers, but there are discrepancies between tubes of different manufacturers as to the exact chemical content of the phosphor, the activators used, the persistence, and the exact fluorescent and phosphorescent colors. An effort is now being made toward phosphor standardization.

Cascade Screens

In the past, production of cathode-ray tubes usually involved only a single phosphor material for the screen. Today many screens are still being manufactured using a single phosphor, but a new process has also been developed for certain applications generally utilizing two phosphors, one coated on top of the other. The screen formed in this manner is referred to as a *cascade* screen.

The primary purpose of such cascade screens is to produce an over-all screen material that has a very long persistence with a high efficiency. One phosphor is placed on the inside glass face of the cathode-ray tube and the other phosphor is coated over this first one. The phosphor nearest the face of the tube is the one having a long-persistence characteristic, while the other usually has a short-persistence characteristic that will not interfere with the phosphorescent color of the first phosphor. Such screens have found extensive use in radar applications during the war where very long-persistence screens were desired because of the extremely long sweep periods used.

The production of light output from the long-persistence phosphor is actually a combination of two processes. Light output is produced from both screens in a manner similar to that previously described due to bombardment by the electron beam. The short-persistence phosphor produces the greater light output by this method because it is the first one in the path of the electron beam. For the long-persistence screen to produce light output by this method, the electron beam must penetrate completely through the first phosphor and then hit the one lying next to the glass face of the tube. If the electron beam penetrates this long-persistence phosphor at all, it will do so for a very short distance. Consequently, the light output from the long-persistence screen due to electron bombardment is very small. The light output from the first phosphor due to this method is quite high because of the high beam penetration and characteristic of the phosphor used. It is the *light output* from the first phosphor which is primarily used to excite the second (long-persistence) phosphor.

Phosphors may be excited into emitting light energy by means other than electron impact. Electromagnetic waves themselves, and hence light, can also be made to excite phosphors into luminescence. What happens in the case of the cascade screen is that the first phosphor emits light due to electron bombardment which is absorbed by the second phosphor, and in turn, produces its own radiant power output within the light spectrum.

The light output from the long-persistence screen is primarily caused by light absorption and only to a small degree by electron impact. Types P7 and P14 made by many manufacturers are typical cascaded screens. The characteristics of these and other types of screens are covered in a later section of this book.

Beam Penetration

Light output due to electron impact is, as we already know, primarily brought about by the release of radiant energy when electrons which have been knocked out of unfilled shells of atoms fall back into these shells. The amount of light output can be increased by the increase in the beam velocity which, in turn, is brought about by an increase in accelerating potential. The increase in electron beam velocity imparts a greater kinetic energy to the beam and this increased energy helps to produce a greater radiant energy output from the screen material. Actually, however, this is not due to a simple multiplication of secondary electrons, since the number of released electrons per unit penetration of the electron beam is more or less *constant* and is independent of the velocity of the electron beam. The increase in electron-beam velocity (energy) increases the light output chiefly because the beam penetrates the screen material deeper and new areas are, therefore, subject to electron bombardment. The cross-sectional area drawing of Fig. 7-5 will make this clearer. This drawing is greatly exaggerated; the beam penetration is only in the order of a few angstroms deep. Let us assume that the electron beam has a certain kinetic energy and that it penetrates the screen material a certain distance as indicated by the first arrow. Let us further assume that the area of phosphor which produces light output is indicated by section *A*. When viewing the face of the tube in the normal manner, a certain amount of light output is seen due to this initial penetration.

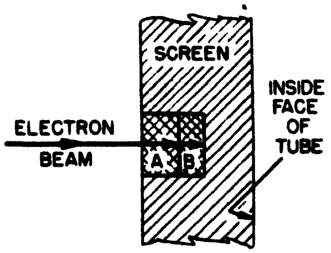

Fig. 7-5.—Penetration of the beam into the screen; the arrow indicates the actual penetration of the beam, while *A* and *B* are the respective fluorescent areas.

Now assume that the accelerating potential of the tube is increased, thereby increasing the velocity and energy of the beam. The greater resulting depth of penetration is indicated by the second arrow and the added amount of phosphor involved in producing light output is indicated by section *B* in the drawing. The higher-velocity beam, in passing through section *A*, does not cause any change in the radiant light output, because the number of electrons released per unit area are independent of the beam velocity. When viewing the tube under the condition of increased beam velocity, the light produced by sections *A* and *B* is effectively *added* to produce a greater light output.

This analysis is greatly simplified, the complete process being much more complex. One point not illustrated by Fig. 7-5 is the relationship of the depth of penetration to the electron-beam velocity. Electron penetrations into *solid* substances are proportional to the square of the kinetic energy of the electron.[2] Consequently, since electron penetration is proportional to the square of the kinetic energy, the depth of penetration is proportional to the *fourth power* of the velocity of the electron beam. Phosphors are not truly solid substances, but for practical purposes, we can assume this fourth-power relation to be a rough approximation. Furthermore, the kinetic energy of the electron beam is directly proportional to the applied accelerating potential so that the penetration of the electron beam is also approximately proportional to the square of the accelerating voltage.

Screen Thickness

The above analysis relative to the beam penetration is strictly true only for a solid substance, and, as we pointed out, phosphors used for screens are not uniformly solid. The screen essentially consists of many minute crystalline particles of various sizes which are not packed tightly in the process of application to the face of the cathode-ray tube. Consequently, many empty spaces exist within the screen material. This means that the electron beam will not always strike the phosphor particles, but may be passing through various empty spaces between the crystalline particles. Other factors complicating the analysis indicate that the electrons that make up the beam do not all have the same velocity as they hit the screen and that every part of the phosphor hit will not necessarily produce light output.

With all these factors in mind, it appears that the optimum thickness of the screen material is dependent upon various conditions and not just upon the accelerating potential of the tube, although this latter factor does play the most important role. If the screen material is too thin, the electron beam may penetrate completely through the phosphor, hit the inside face of the tube, and cause electron burning of the glass. Because of the above considerations, it appears that the screen material should have a greater thickness than that based upon electron penetration into *solid* material. This is exactly what is done in practice, but there is also a limit to how thick the screen may be. This limiting thickness is referred to as the *optimum thickness* of the screen for most efficient light output. If the layer of screen material is made too thick, part of the light emitted by one portion of the screen may be absorbed by another portion without re-emitting any visible radiation.[3] In other words, what generally happens is that the beam only penetrates a small portion of the screen and excites this part into visible radiation. The light rays from this part of the screen may be absorbed by other portions which are neither excited by the beam nor by the light rays from the first part of the screen. In general, the greater the thickness beyond the optimum value, the greater is the reduction in screen efficiency.

The optimum thickness, besides being a function of the type of phosphor used, is also dependent upon the modes of screen excitation. The strength of excitation, which is a direct function of the kinetic energy of the beam, is, as we know, one factor that plays a very important role. However, the duration of excitation is also important. If the excitation is continuous, the screen would normally have to be somewhat thicker than for discontinuous or pulsed operation. In other words, whether the excitation is continuous or intermittent is an important factor to be considered. Since the accelerating voltage has the greatest single effect upon the electron-beam penetration of the screen, the optimum screen thickness is chosen primarily in accordance with the accelerating voltage that is to be used. However, for more accurate determination of optimum screen thickness, all conditions of excitation to which the tube may be subject should be taken into account.

Heat Generation

When the electron beam strikes the screen, not all of the kinetic energy is converted into light energy or into the production of secondary electrons. Heat energy is also produced. This heat energy can be very detrimental to the production of light output from the screen. The efficiency of light emission from most screens is dependent upon the temperature of the phosphor, among other factors. It has been found that, as the temperature rises, the efficiency of the screen phosphor decreases.

If the electron beam is kept moving and is deflected all around the screen, the screen will be maintained at a low average temperature. However, if the beam is concentrated at a small part of the screen, the heat generated is then confined and the temperature of that area rises. This can occur even if the beam is deflected, as in tracing out the same sine wave continuously over and over again, but the temperature will rise even more when the beam remains stationary and causes only a spot of light to appear.

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[2] Soller, T., Starr, M. A., Valley, G. E., "Cathode-Ray Tube Displays," McGraw Hill Book Co., New York, N. Y., 1948; p. 628.

[3] Feldt, R., "Luminescent screens for cathode-ray oscillography," *The Oscillographer*, vol. 11, April-June, 1949.

Spot Burn

If the electron beam continues to strike the same area for any prolonged length of time, the higher temperature produced may cause permanent loss of efficiency from that part of the screen. The screen may actually become burned due to the high temperatures, resulting in the so-called *spot* burn. To prevent damage to the screen material, the beam should not be allowed to remain stationary in any one spot, or even to be deflected over the same path for any long period of time. Decreasing the intensity of the beam under these circumstances will help prevent screen burns.

Spot Size

In previous chapters, we discussed the factors that influence the size of the spot. These factors dealt primarily with the electron gun and the general laws of optics. As far as the screen material itself is concerned, two principal factors which affect the size of the spot are the thickness of the screen used and the interaction among electrons during bombardment.

An increase in accelerating potential will reduce the spot size. The thickness of most screens, as mentioned previously, is principally determined on the basis of the required accelerating voltage. If a screen designed for a high accelerating voltage is used with a low accelerating voltage, the spot size will increase. A cascaded screen, therefore, will show a larger spot than a single-layer screen due to the condition of operation of such screens. Furthermore, the electrons themselves, being negative in character, have a repelling effect on each other which also increases the spot size.

Screen Potential

We have mentioned before that the screen phosphor becomes charged due to electron bombardment. This charging phenomenon makes the screen exhibit a potential. Since the screen material is uneven, that is, it is not smooth, and causes sections to become charged to different values, the screen potential would seem to increase, especially if the phosphor is unevenly spread on the face of the tube. This is true to some degree, but, as mentioned before, slow-moving secondary electrons are attracted back to certain portions of the screen, helping to reduce the screen potential to a low value.

Halo

Several things can happen to the light output produced by the screen, one of these being the so-called *halo effect,* which results in a considerable loss of picture contrast. Let us consider this effect in detail, referring to Fig. 7-6A. The light spot X on the screen emits light rays in all directions. Ray a, which is perpendicular to the screen, passes normally through the glass face of the tube without any change. Rays b and c, initially forming a small angle with the perpendicular to the screen, are refracted at the glass according to the well-known laws of optics; that is, they are tilted away from the perpendicular in passing out from the glass face. At a certain so-called *critical angle,* this tilt becomes so great that the ray of light is no longer able to emerge from the glass face, because it would be parallel to it. At this point, the light is *totally reflected* back toward the screens. This is shown in Fig. 7-6A for rays d and e emitted at the critical angle θ. These rays, after being reflected toward the screen, reactivate

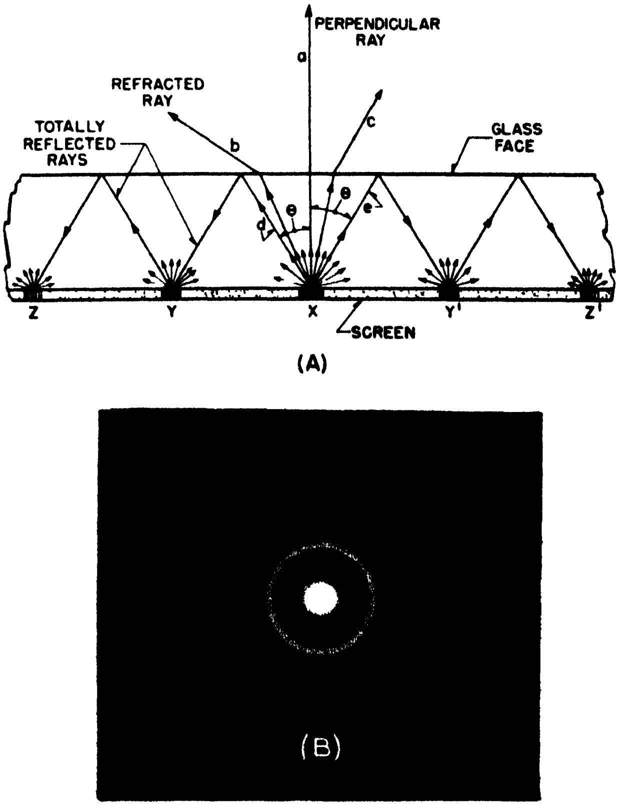

Fig. 7-6.—(A) The halation effect produced by totally reflected rays returning from the outer glass face to the screen, resulting in secondary fluorescence, while (B) is an actual photograph taken from the screen of a cathode-ray tube.

the phosphor and produce further light output at points *Y* and *Y'* of the screen.

It is obvious that this additional light will interfere with the normal shadows and lights of the picture and, therefore, reduce contrast. *Y* and *Y'* are only two points on the cross section of the tube face shown on the screen; actually, a whole ring of light-producing points concentric with light spot *X* is formed. This ring is called the *halo*. However, usually more than one concentric ring is produced; this occurs in the following manner. The screen, being a nonuniform (nonhomogeneous) crystalline material, is not everywhere in optical contact with the glass surface. Those parts of the screen which are in contact with the glass face at points *Y* and *Y'* will be illuminated by the totally reflected rays *d* and *e* and produce light output. Those parts of the screen, however, which are not in optical contact with the glass, cannot produce light. These portions of rays *d* and *e* are, therefore, again totally reflected at *Y* and *Y'* toward the outer glass surface. Here, they are, in turn, totally reflected toward points *Z* and *Z'* of the screen, forming another halo ring of weaker intensity. In this way several rings may be produced, each progressively weaker in intensity. Fig. 7-6B is an actual photograph, showing the light spot in the center and two concentric halo rings.

Characteristic Curves

In order to indicate the performance of phosphor screens relative to screen efficiency, radiant energy distribution in the visible spectrum, persistence qualities, and writing speeds, various characteristic curves have been devised. In this section, we will study such curves so that we can become familiar with practical aspects.

Eye Visibility Curve

The limitations of the visible spectrum arise out of the inability of the human eye to respond equally to different electromagnetic radiations. In Fig. 7-4 the visible spectrum was indicated to extend approximately from wavelength of 4,000 Å to 7,000 Å. In reality, there are no exact limits to this spectrum because not all eyes respond the same. The limits chosen are considered to be the average limitations of the eye's response. A knowledge of the eye's sensitivity to light is necessary in order to understand the spectral characteristic curves of the different phosphors.

The eye *does not* have the same response to light of different frequencies. If we were to look at different monochromatic (single-frequency) lights of the same intensity from one end of the visible spectrum to the other, the relative *brightness* of each color would seem to increase to a maximum point gradually and then decrease again. This sensitivity of the human eye is graphically represented by the curve of Fig. 7-7, commonly called the *eye visibility curve* or the *eye sensitivity curve*. The colors of the spectrum are indicated on top of the graphs for the sake of illustration.

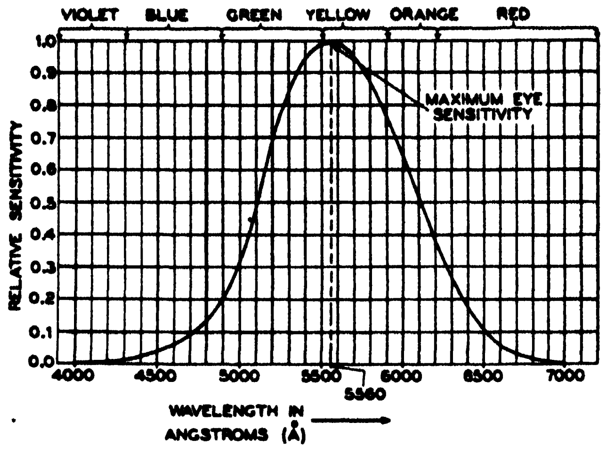

Fig. 7-7.—Relative sensitivity of the eye to different wavelengths of the visible spectrum; maximum sensitivity occurs approximately at 5,560 Å.

This curve is drawn for the average light-adapted eye and it is seen that the maximum sensitivity occurs at a wavelength of approximately 5,560 Å. This means that a greenish-yellow light appears the brightest. The intensity of the different wavelengths for this curve is considered to be average—that is, neither too weak nor too bright. When the intensity of the light is increased or decreased, the visibility curve will change, with the maximum sensitivity occurring at a different wavelength. For example, at weak intensities, the eye's response is more sensitive to light in the lower wavelength region with the maximum sensitivity being around the green or blue region of the spectrum. Whatever the case may be, it should be remembered that the eye does not have the same sensitivity to all colors in the visible spectrum and that its maximum sensitivity can vary somewhat according to the intensity of the light being observed and the physical characteristics of the particular eye.

Spectral-Energy Distribution

The energy radiated from a phosphor screen is not at one frequency in the visible spectrum but covers a range of wavelengths. Also, the intensity of the radiant energy is not the same for the different wavelengths radiated. In order to indicate these variations for the different phosphors, special curves have been devised which show the intensity versus wavelength characteristic for each phosphor. These curves are known by such terms as *spectral-energy-distribution* curves or *spectral-energy emission characteristic* or just *spectral characteristic.*

In Fig. 7-8 appears a typical spectral-characteristic curve. The abscissa or X-axis of the graph represents wavelengths in terms of Angstrom units. The ordinate or Y-axis represents the radiant energy output from the phosphor. This part of the graph is usually expressed in terms of relative values with the top or peak part of the curve given a maximum intensity value of 100 (as used in Fig. 7-8) or a maximum value of unity.

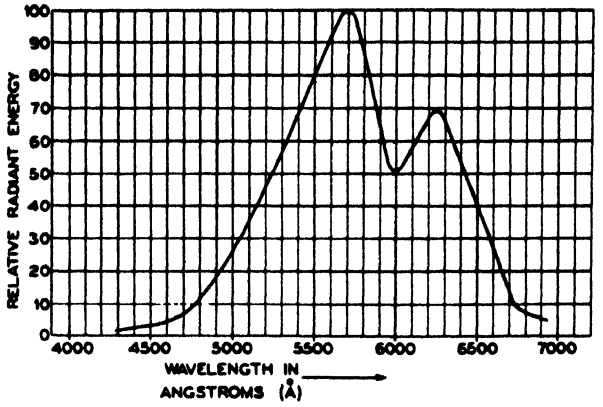

Fig. 7-8.—Typical spectral-characteristic curve of a phosphor, showing two radiant energy peaks at two different wavelengths.

The ordinate, besides being called the relative radiant energy, is also commonly known as relative energy or relative intensity.

The exact shape of the curve will vary for the different types of phosphors tested. The method of measurement used to obtain these curves is unimportant to us, but what is important is to understand that such curves do exist and from them the amount of radiant energy at different colors (i.e., wavelengths) can be seen easily. For example, from the curve of Fig. 7-8, the maximum radiant energy output occurs at 5,700 Å, which is in the yellow region of the spectrum. It can also be seen that the sensitivity in the orange region (above 6,000 Å) and green to blue regions (5,200 Å and below) has decreased to less than 50 per cent of the maximum sensitivity. Such curves are plotted under a specific set of conditions of cathode-ray-tube operation, the accelerating potential and second-anode or intensifier-electrode current per unit screen area being the most important factors. The interesting thing about the curve shown here is that it has two peaks. Such curves are usually the result of using two different phosphors for the same screen material.

Visual Spectral Characteristic

The curve of Fig. 7-8 tells us the relative amount of radiant energy output at different wavelengths. Another interesting and important curve refers to the luminous output of the screen at different wavelengths as seen by the human eye and is generally referred to as the *visual spectral characteristic* or visual response curve. It is a plot of luminous energy output (usually on a relative basis) with respect to wavelength corrected to the eye's sensitivity curve. The ordinates or Y-axis values of such curves for each specific wavelength are obtained by multiplying the values of the relative radiant energy from the spectral-energy-characteristic curve by the relative sensitivity from the eye's visibility curve.

As a typical example, let us use the spectral characteristic of Fig. 7-8 in conjunction with the eye visibility curve of Fig. 7-7 to obtain this new curve. The resulting visual spectral characteristic is shown in Fig. 7-9. To indicate how this curve was plotted, let us examine the relative radiant energy of Fig. 7-8 and the relative sensitivity of Fig. 7-7 at three different wavelengths, namely, at 5,000 Å, 5,500 Å, and 6,000 Å. From Fig. 7-7, the relative eye sensitivity at these three wavelengths is seen to be approximately equal to 32, 98, and 64 per cent of the maximum sensitivity, respectively. From Fig. 7-8, the relative radiant energy at these three frequencies is found to have values of 26, 80, and 51, respectively. By multiplying 0.32 by 26, 0.98 by 80, and 0.64 by 51, the ordinates or Y-axis values for the visual spectral characteristic of Fig. 7-9 will be found. The products of these values are 8.3, 78.4, and 32.6, respectively, and are thus indicated in the graph of Fig. 7-9.

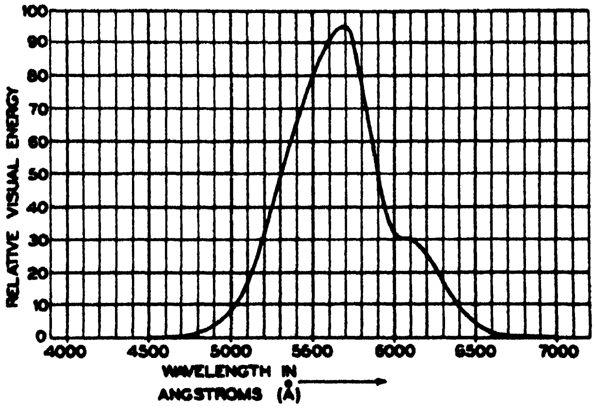

Fig. 7-9.—The resultant *visual* spectral characteristic of the phosphor shown in Fig. 7-8.

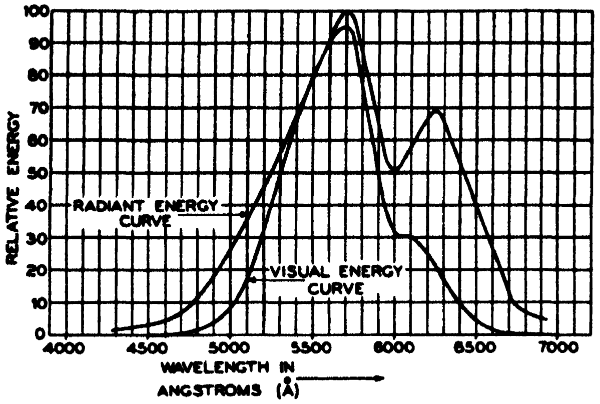

Fig. 7-10.—The difference in area between the graphs of Figs. 7-8 and 7-9 provides an indication of the visual efficiency of the phosphor.

If the graph of Fig. 7-9 is superimposed over the graph of Fig. 7-8, the difference in area enclosed by both curves will be a relative over-all indication of the visual efficiency of the phosphor. This efficiency is related to the amount of luminous or visual light output as compared to the total amount of radiant energy available.[4] The superimposed graphs are illustrated in Fig. 7-10. The ratio of the area under the visible energy curve (the smaller area) to that under the radiant energy curve in terms of percentage tells us how much of the total radiant energy is utilized in the production of light output as far as the human eye is concerned. If the ratio is, say, 0.70, it means that only 30 per cent of the radiant energy is not "seen" or "utilized" by the eye and that 70 per cent is useful. Such a value is considered highly efficient for cathode-ray-tube screens.

Screen Efficiency

The efficiency of the screen in producing light output in accordance with the applied *accelerating potential* is another basic characteristic of cathode-ray-tube screens. We already mentioned that the amount of light output from screens is dependent upon the accelerating voltage as one principal factor. According to the type of phosphor used as the screen, there are upper and lower limits of useful accelerating potential. Below the lower limit, there will not be any light output and above the upper limit, the light output will not increase any more, remaining more or less constant.

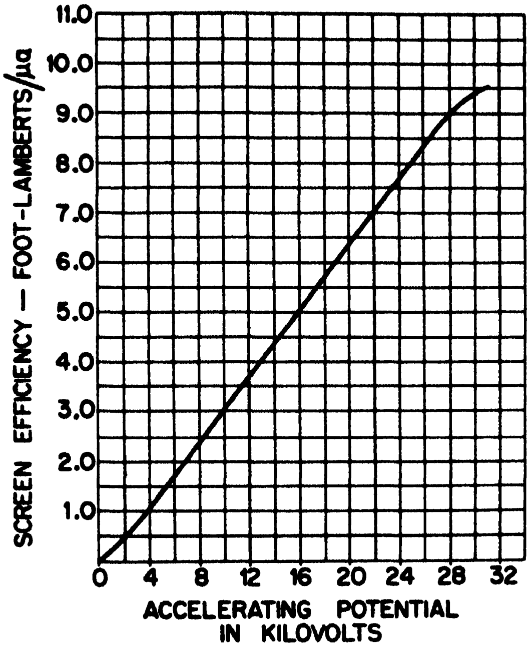

Fig. 7-11.—The screen efficiency of a typical phosphor in terms of foot-lamberts/μa versus the accelerating potential.

The screen efficiency graph is not shown normally in terms of percentage, but rather in terms of light output in foot-lamberts. For example, let us examine the simple curve of Fig. 7-11. This curve represents the light output *per microampere* of beam current with respect to the applied accelerating voltage at one specific value of beam current. You will notice that at 16 kilovolts, the screen efficiency is found to be about 50 foot-lamberts per microampere. This means that if the beam current for this curve is 10 microamperes, the total light output is 50 foot-lamberts.

Most screen efficiency graphs for any one phosphor contain more than one curve, each for a different beam current. For any one particular set of bias conditions, the beam current is essentially the same, so that for different beam currents only the bias voltage of the cathode-ray tube is primarily changed. The circumstances under which these curves were obtained are also quite important. Information as to the exact size of the section of the screen (called the raster) used for the light output measurements, the type of instrument used, and how it is calibrated, should be given for each curve.

Two principal types of screen efficiency curves exist. The light output from one is measured with a type of light meter that does not take into account the sensitivity of the human eye. Such types of curves are normally referred to simply as *average screen* efficiency curves. The other type of curve is called the *average visual screen efficiency* and is obtained by means of a special-type light meter that is corrected for the sensitivity of the human eye. In studying the characteristics of the different screens in the next section, both types of curves will be illustrated.

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[4] Perkins, T. B., Kaufman, H. W., "Luminescent materials for cathode-ray tubes," *Proc. I.R.E.*, vol. 23, p. 1331, November, 1935.

Persistence Characteristic

The brightness of the screen after the excitation has been removed, that is, during phosphorescence, generally decreases in a logarithmic or exponential fashion. A plot of the brightness also referred to as the brilliance versus the decay time of the phosphorescence is termed the *persistence characteristic* of the screen. The brightness, normally given in relative units, is plotted as the ordinate or *Y*-axis on a logarithmic scale, while the decay time is plotted on a linear scale along the *X*-axis.

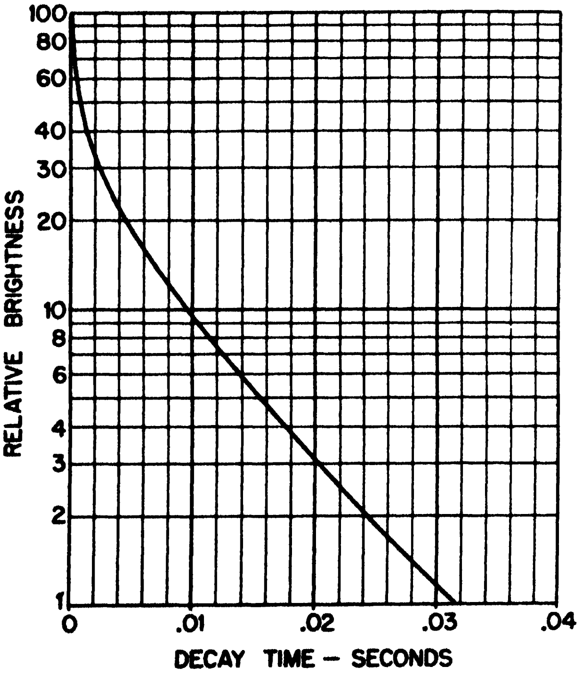

Fig. 7-12.—A typical persistence characteristic for a phosphor of medium persistence.

In Fig. 7-12 appears a typical persistence characteristic. In this curve, decay time is indicated in seconds, but it is also sometimes shown in units of milliseconds or microseconds. The brightness scale, here given in relative units as a percentage of the maximum, can also be shown in units of brightness, such as footlamberts. The curve illustrated applies to a phosphor of medium persistence. This curve tells us that after about 0.032 seconds (32 milliseconds) the relative brightness will have decreased to approximately one per cent of the maximum brightness.

Average Characteristic Curves

The luminous intensity per unit area of the phosphor will vary in accordance with the accelerating potential and also the second-anode or intensifier-anode current. When the luminous intensity is plotted against this

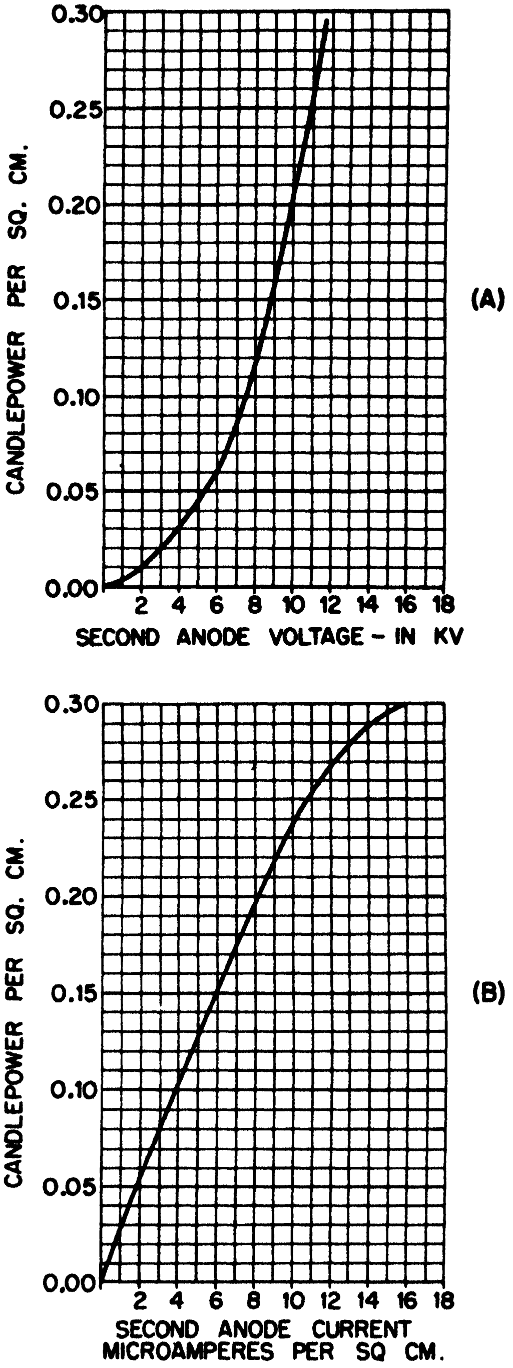

Fig. 7-13.—(A) Average luminous intensity per unit area of a typical phosphor; the intensity in candlepower per square centimeter is shown for different second-anode potentials, while (B) indicates the variation in luminous intensity with the second-anode current as abscissa.

current or voltage, the type of curves that result are generally referred to as average phosphor characteristics. The luminous intensity is normally chosen as the ordinate and calibrated in terms of candlepower per unit screen area. The abscissa is usually given either in terms of second-anode voltage or current.

Two typical graphs appear in Fig. 7-13. In that at (A), the abscissa is in terms of second-anode voltage, and in that at (B), it is in terms of second-anode current. In the curve of graph (A), the *current* per square centimeter (the same unit area as used for the Y-axis designation) is held constant, and for the curve of graph (B), the second-anode voltage is kept constant. In practice, a family of curves is usually given in each type of graph, with the constant current or voltage condition indicated for each curve. It is seen that the curve at (A) follows approximately the square-law relationship discussed earlier in the theoretical part of the chapter.

Highlight Brightness

The average brightness of a screen for a given second- or third-anode current will vary according to the size of the raster from which it is measured and the value of accelerating potential used. Some screen manufacturers have characteristic curves indicating the brightness in terms of such units as foot-lamberts versus anode current for a number of different raster sizes from which the measurements are made. Such curves are often referred to as ***highlight-brightness curves.***

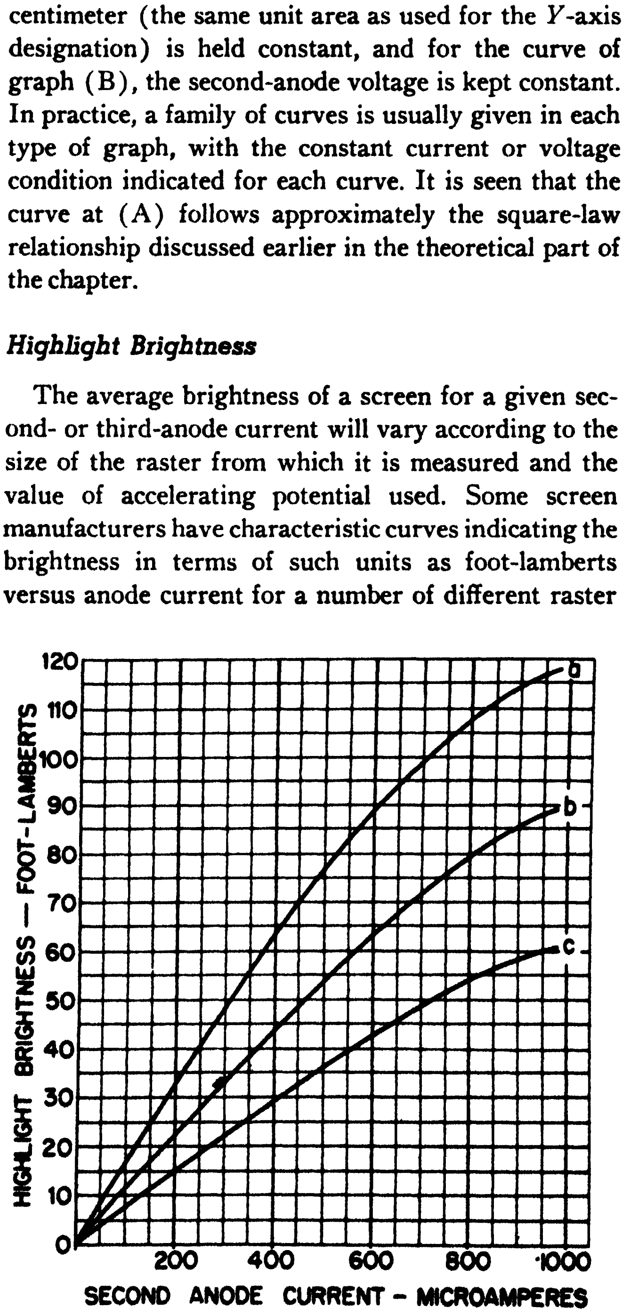

Fig. 7-14.—Variation in average highlight brightness with second-anode current for different raster sizes; curves *a*, *b*, and *c* are in the order of increasing raster size.

In the highlight-brightness graph of Fig. 7-14, three curves are drawn. Each curve is for a different size raster. The brightness decreases with increase in raster size so that curves *a*, *b*, and *c* are for rasters of increasing size. Such graphs are often used for screens employed in television work where the raster size has a ratio of 4 to 3, the same as the television aspect ratio. As an example, for a second-anode current of 600 microamperes, the highlight brightness for curves *a*, *b*, and *c* of Fig. 7-14 decreases from 88 to 63 to 42 foot-lamberts.

Writing-Speed Characteristics

Both for visual observation and photographic work, the writing speed of different screens is very important, especially for the observation of transient phenomena and radar displays, such as used in PPI. In this section we will discuss writing-speed characteristics pertaining to visual observation.

For a standard, a sinusoidal signal will be used. The writing speed of the beam is directly related to the frequency and amplitude of the applied signal. The speed of the electron beam, that is, the beam spot, due to a sinusoidal applied signal, is not constant, but varies, and the greatest speed occurs when the sine wave passes through its zero voltage points (i.e., its baseline). The general equation for the velocity v of a sine wave is as follows[5]

$$v = \omega A \cos \omega t \tag{7-1}$$

where A is the peak amplitude of the sine wave and ω the angular velocity equal to 2π × frequency. Since the maximum value the cosine can have is unity (at 0° or 180°), the magnitude of the maximum velocity V_m is

$$V_m = \omega A = 2\pi f A \tag{7-2}$$

where f is the frequency.

The writing speed of the spot is the vector sum of the horizontal sweep writing speed and the vertical maximum signal writing speed V_m. However, if the velocity of the linear time-base signal (i.e., the horizontal sweep) is much smaller than that of the sinusoidal input signal, at least in the ratio of 1 to 10, the

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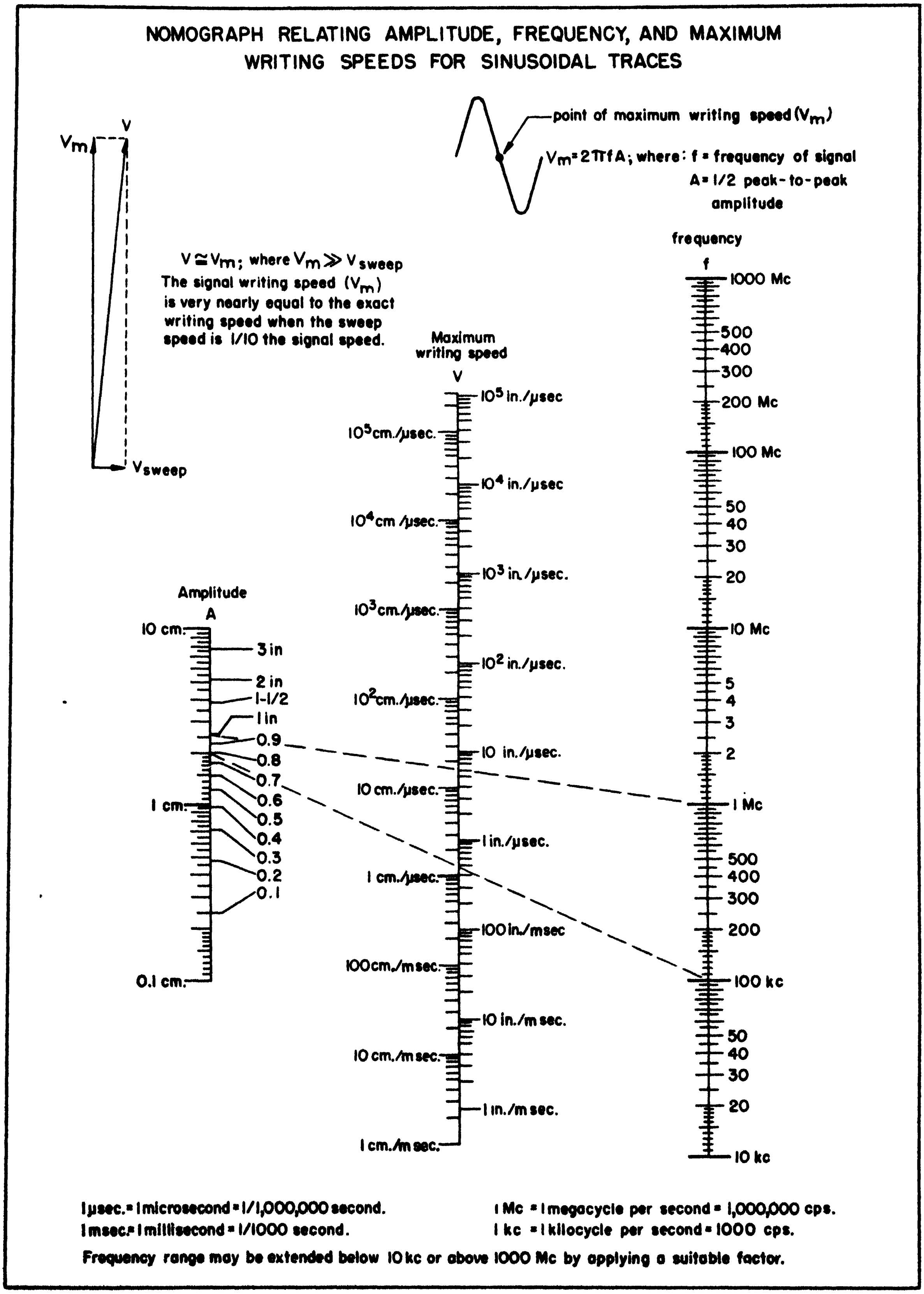
[5] A sine wave is simple harmonic motion and is given by the equation $y = A \sin \omega t$ (where the displacement y in this case is in the vertical direction). The velocity of simple harmonic motion or the time rate of change of the displacement is the derivative of displacement, namely dy/dt and hence the velocity becomes

$$v = \frac{dy}{dt} = \omega A \cos \omega t.$$



1µsec. = 1 microsecond = 1/1,000,000 second.
1 msec. = 1 millisecond = 1/1000 second.

1 Mc = 1 megacycle per second = 1,000,000 cps.
1 kc = 1 kilocycle per second = 1000 cps.

Frequency range may be extended below 10 kc or above 1000 Mc by applying a suitable factor.

Courtesy Du Mont Labs.

Fig. 7-15.—Alignment chart showing the relation between maximum writing speeds of sinusoidal traces, frequency, and amplitude in metric and English units of measurement.

actual writing speed of the beam spot is considered to be essentially equal to the maximum writing speed of the sine wave. Let us calculate a few writing speeds on this basis.

If a 1-Mc sine wave has a peak-to-peak amplitude of 2 inches ($A = 1$ inch) the maximum writing speed V_m will be

$$\begin{aligned} V_m &= 2\pi \times 10^6 \times 1 \\ &= 6.28 \times 10^6 \text{ inches per second} \\ &\text{or } 6.28 \text{ inches per microsecond.} \end{aligned}$$

If a 100-kc sine wave has a peak-to-peak amplitude of 4 cm ($A = 2$ cm), the maximum writing speed V_m will be

$$\begin{aligned} V_m &= 2\pi \times 100 \times 10^3 \times 2 \\ &= 1.256 \times 10^6 \text{ cm per second} \\ &\text{or } 1.256 \text{ cm per microsecond.} \end{aligned}$$

In Fig. 7-15 appears a nomograph which relates the maximum writing speed (here assumed to be equal to the actual writing speed V) to the frequency and amplitude of the sinusoidal signal. This chart takes into account signal amplitude measurements in either the English or metrical systems. If the peak amplitude of the sine wave is measured in metric units, the left sides of the A and V columns are to be used, and if the measurement is in English units, the right sides of the A and V columns are to be used. All scales shown are logarithmic. This chart is used in a manner similar to most nomographs. With the frequency and peak amplitude known, it is a simple matter to determine the writing speed. Using a straight edge, line up the frequency of column f with the peak amplitude of column A and at the point of intersection in the V column, the maximum writing speed can be read off. The two dashed lines in Fig. 7-15 indicate the method of determining the writing speed for the two previous examples where $f = 1$ Mc, $A = 2$ inches, and $f = 100$ kc, $A = 4$ cm. From the V column of Fig. 7-15, the writing speeds are immediately seen from the intersections. Note how these points check with the calculated values. The nomograph can be used to find f, V, or A, as long as at least two of the quantities are known.

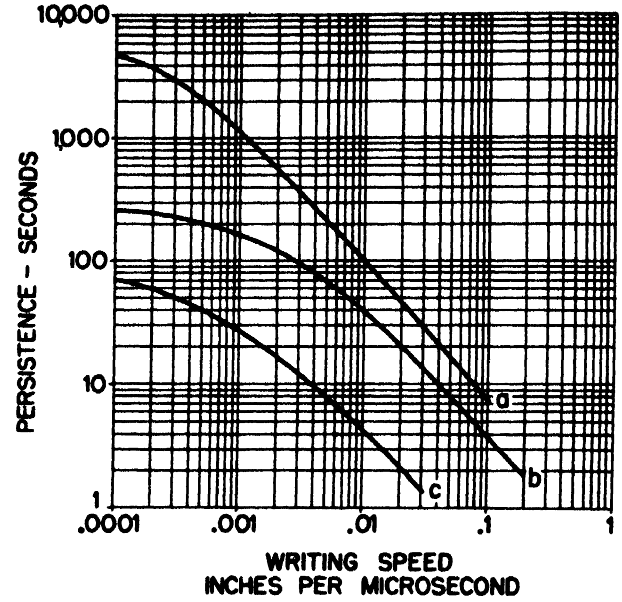


Fig. 7-16.—Persistence versus maximum writing speed characteristics for different constant accelerating potentials; curves *a* through *c* are in the order of decreasing potential.

There are two principal types of graphs in which the (maximum) writing speed is utilized. In one type, brightness of the screen is plotted against writing speeds, and in the other type, persistence is plotted against writing speeds. The latter type of graph is used more often, a typical example of which appears in Fig. 7-16. This graph, which is a log-log plot, is for a long-persistence screen. It is usually plotted on the basis of single sinusoidal transients at some specific amplitude and beam current and is useful for only one type of phosphor. The conditions under which each of the three curves was plotted are the same with the exception that each represents a different accelerating potential with the potential decreasing from curve *a* to *c*. The light conditions, upon which such measurements are based, are usually also given.

Although the graph is originally plotted for a specific signal amplitude, it can also be used for other signal amplitudes. To do this, the writing speed must first be determined from the nomograph of Fig. 7-15, or from a single calculation of $V_m = 2\pi fA$. Once the writing speed is known, the persistence time can then be established. For instance, for a writing speed of 0.01 inches per microsecond, we find that, for curve *b*, the persistence is approximately 40 seconds. This means that, for the signal input in question, it will be 40 seconds before the afterglow or phosphorescence will no longer be seen by the average eye. From this curve we see that, for slower writing speeds, the persistence period is longer.

COMMERCIAL CATHODE-RAY-TUBE SCREENS

Although there is no final accepted standard relative to cathode-ray-tube screens, there is nevertheless some effort made toward standardization by the manufacturers themselves. For the same *P* number, you will

TABLE I PHOSPHOR RATINGS

Phosphor No.	*Fluorescence*	*Phosphorescence*	*Persistence*	*Applications*
P1	Green	Green	Medium	General oscilloscope use, such as for periodic waveform observation.
P2	Blue-Green	Green or Yellow-Green	Long	General oscilloscope use, primarily suited for slow and medium transient phenomena.
P3	Yellow	Yellow	Medium	General oscilloscope use, for periodic waveform observation.
P4	White	White	Medium	Principal use is for television picture tubes.
P5	Blue	Blue	Very Short	Principal use in oscillography where high-speed photographic recording is required.
P6	White	White	Short	Used primarily for color television.
P7	Blue-White	Greenish-Yellow or Yellow	Very Long	Used for observation of slow and medium transient phenomena. First used in radar.
P11	Blue	Blue	Short	Principal use is for photography, but for lower speeds and higher efficiency recordings than the P5 phosphor.
P12	Orange	Orange	Long	Primarily used for radar.
P14	Purple or White	Orange	Long	At first used for radar, but also applicable to slow and medium transient phenomena, shorter persistence than P7 screen.
P15	Blue-Green and Ultraviolet	Blue-Green	Very Short	Specifically suitable for high-speed scanning requirements, such as flying spot scanning, because it has a very short near-ultraviolet fluorescence.

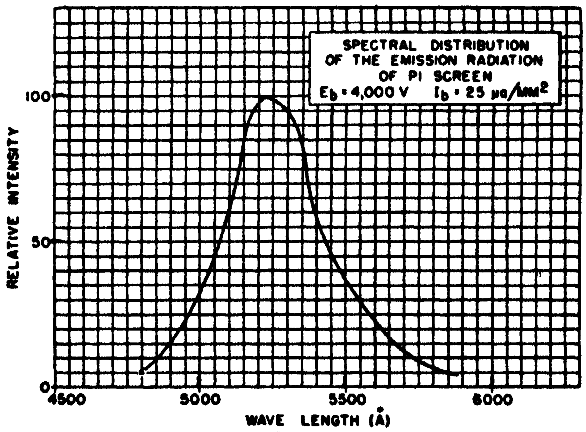
find that many manufacturers classify persistence the same way into short, medium, or long persistence, and list perhaps the same fluorescent color, but the exact persistence time, the chemicals used for the phosphor, the spectral characteristics, and other factors may differ. The information on commercial cathode-ray-tube screens to be included in this section is chiefly based upon specific screens manufactured by two principal tube manufacturers, namely by Allen B. Du Mont Laboratories, Inc., and by Radio Corporation of America.

Phosphor Groupings

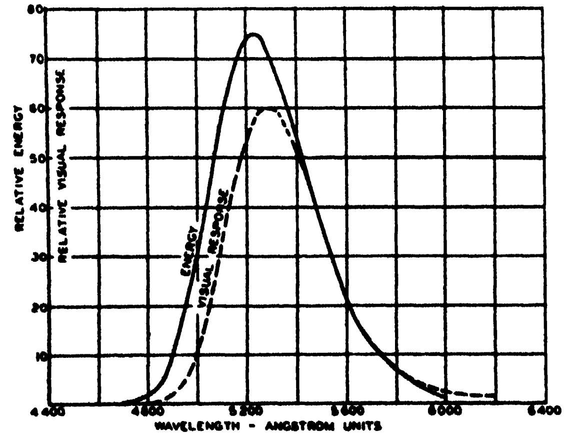
Phosphors for excitation by cathode rays are today classified into 15 general groupings, from P1 to P15. Some of the phosphors are special types designed during the last war primarily for radar applications, among which are the P7, P8, P9, P10, P12, and P13 phosphors. Some of these phosphors are no longer in use. *Available* information on most phosphors, including the fluorescent and phosphorescent color, the persistence rating, and the principal applications, are given in Table I.

The chemical composition of these phosphors has not been listed here, since it is of little practical use and at the date of writing is somewhat controversial. Different manufacturers list varying compounds, of which some are obsolete and no longer manufactured, while the exact composition of others has not been made available at all.

This chart is an over-all cross section of the phosphors as listed by leading manufacturers; an examination of this table will show that the manufacturers



(A) *(A) Courtesy Du Mont Labs.*



(B) *(B) Courtesy Proc. I.R.E.*

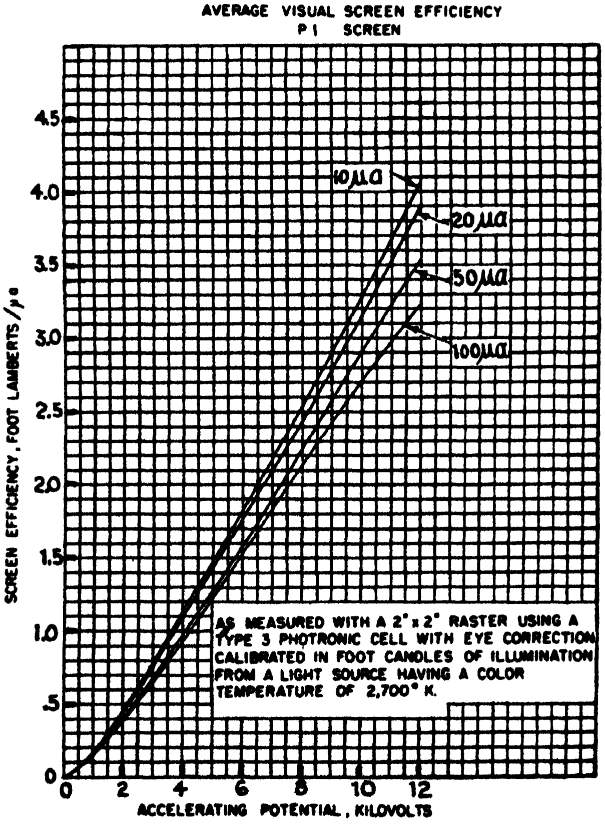
Fig. 7-17.—The spectral characteristic of a typical P1 phosphor is shown in (A), while (B) contrasts the spectral characteristic of a P1 phosphor with a visual spectral characteristic (corrected for the eye's response).

themselves do not differ too much as to the fundamental characteristics and uses. In this section, we will study the characteristic curves of those phosphors used in commercial cathode-ray tubes, including the P-, P2, P4, P5, P7, P11, and P15 phosphors. This chart can be used for quick comparison between the different phosphors. The activator used, if any, is not listed in the chart.

Phosphor P1

This phosphor is one of the most commonly employed today for cathode-ray-tube screens. It is used for general oscillographic work and is considered to have a high visual efficiency. Most oscilloscopes that are sold today normally come equipped with cathode-ray tubes containing P1 screens. In Fig. 7-17A appears the spectral characteristic of a typical P1 phosphor. The conditions under which this curve was taken are indicated in the figure. The maximum intensity occurs at about 5,225 Å, which is approximately in the center of the green region of the visible spectrum; see Fig. 7-4. Since the green region extends from about 4,900 Å to 5,500 Å, it is seen from Fig. 7-17A that most of the spectral characteristics of the P1 screen cover this region. In Fig. 7-17B appears another spectral characteristic of a typical P1 phosphor. This graph includes the visual spectral characteristic as well as the radiant energy characteristic. Due to the high ratio of the areas under each curve, the P1 phosphor is seen to have a high luminous or visual efficiency.

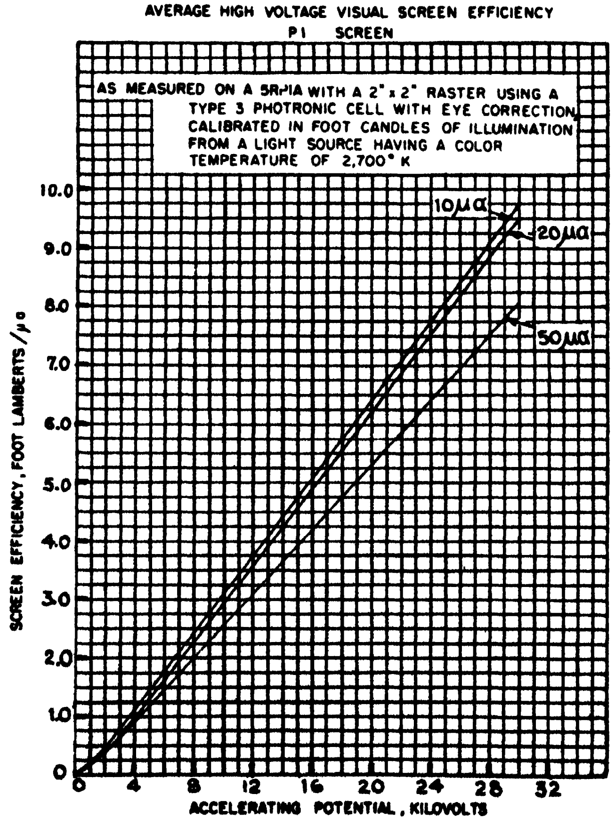
The average visual screen efficiency of a typical P1 phosphor for accelerating potentials of 12 kilovolts and less is illustrated in Fig. 7-18. The conditions under which measurements were made for this graph are indicated on the drawing. Four individual curves are shown, each for a different beam-current rating. For the same accelerating potential, the screen efficiency decreases as the beam current increases. For example, at 10 kilovolts, the screen efficiency for the 10-μa curve is about 3.3 foot-lamberts/μa and for the 100-μa curve, the screen efficiency is only approximately 2.7 foot-lamberts/μa. The amount of light output, however,



AS MEASURED WITH A 2" x 2" RASTER USING A TYPE 3 PHOTRONIC CELL WITH EYE CORRECTION CALIBRATED IN FOOT CANDLES OF ILLUMINATION FROM A LIGHT SOURCE HAVING A COLOR TEMPERATURE OF 2,700° K.

Courtesy Du Mont Labs.

Fig. 7-18.—Average visual screen efficiency characteristic of a P1 phosphor.

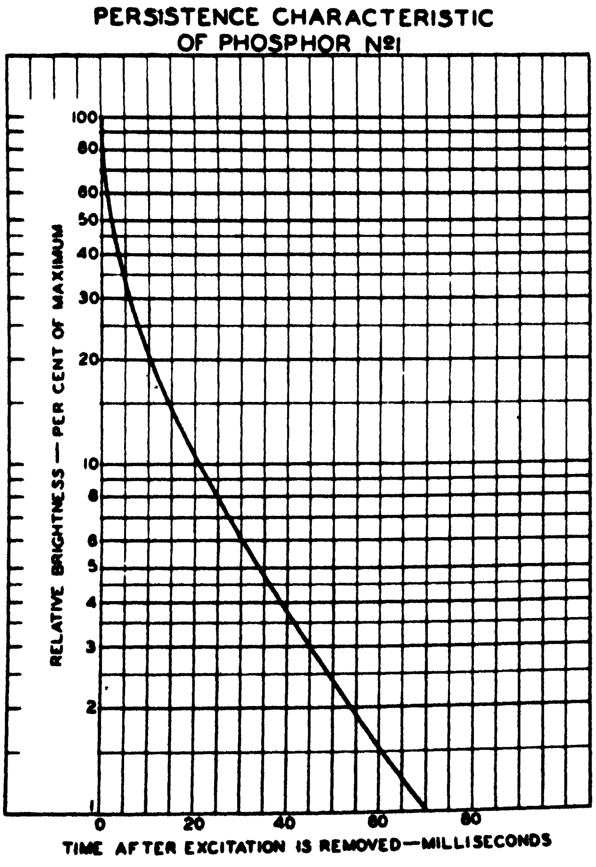


AS MEASURED ON A 5RP1A WITH A 2" x 2" RASTER USING A TYPE 3 PHOTRONIC CELL WITH EYE CORRECTION CALIBRATED IN FOOT CANDLES OF ILLUMINATION FROM A LIGHT SOURCE HAVING A COLOR TEMPERATURE OF 2,700° K

Courtesy Du Mont Labs.

Fig. 7-19.—Average visual screen efficiency of a P1 phosphor for high accelerating potentials.

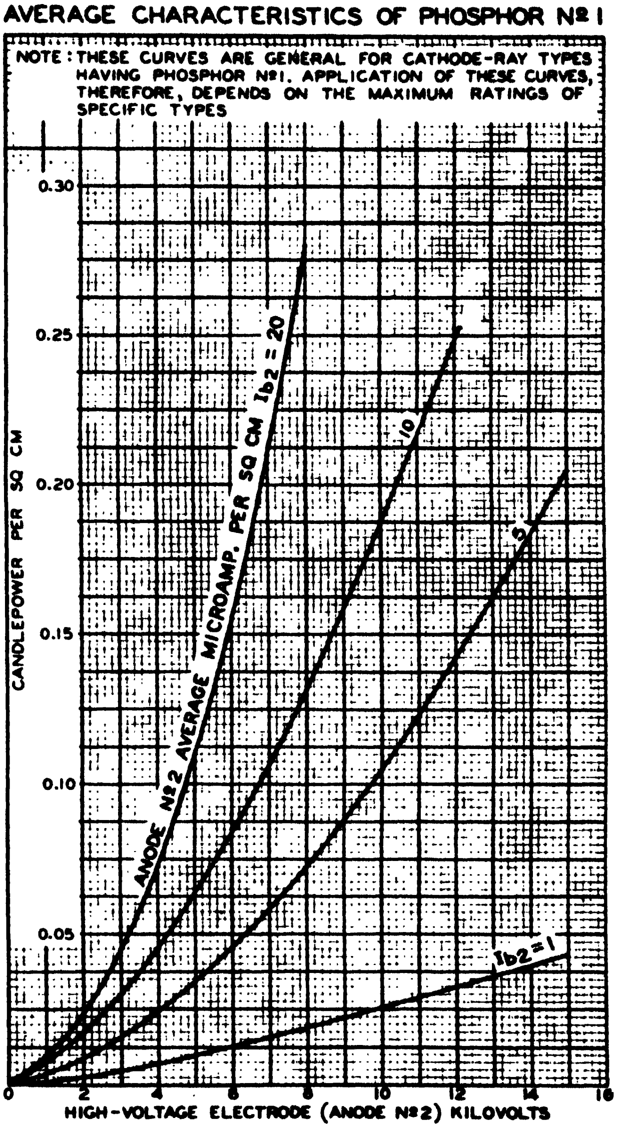
increases. For the 10-μa curve, the output is 33 foot-lamberts, whereas for the 100-μa curve, it is 270 foot-lamberts. The increase in light output is not in the same ratio as the increase in beam current. The graph of Fig. 7-19 is also a visual screen efficiency characteristic for phosphor P1, but it includes much higher accelerating voltages. Only the 10-μa, 20-μa, and 50-μa curves are included in this graph. From these curves, it is seen that the efficiency characteristic of this screen is still quite linear even at higher voltages and for a fair range of beam current.



Courtesy RCA

Fig. 7-20.—The persistence characteristic of a typical P1 phosphor.

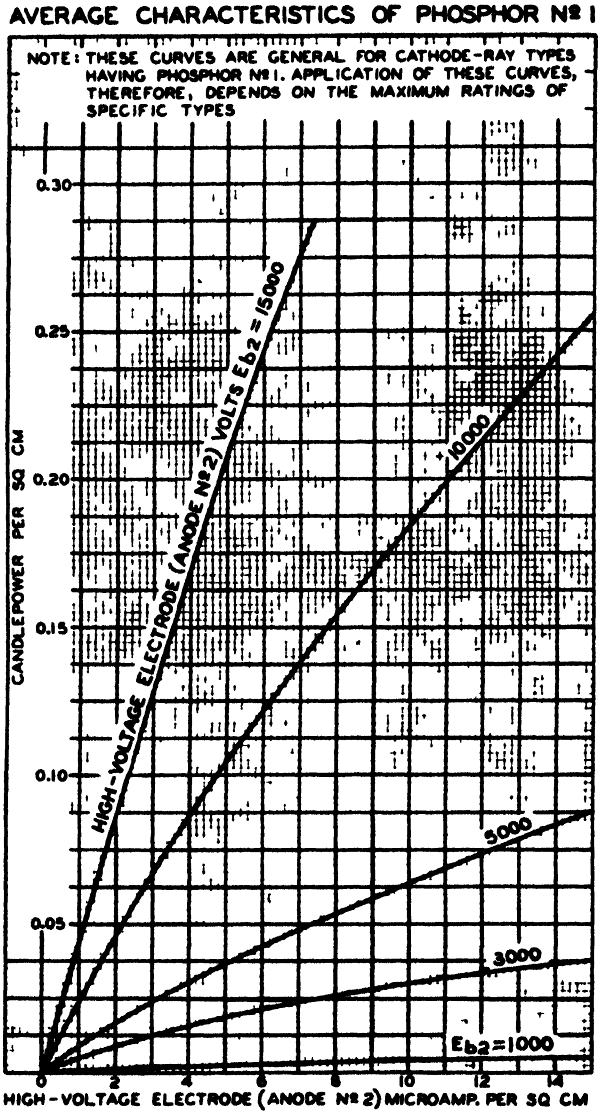
The persistence characteristic of a typical P1 phosphor appears in Fig. 7-20. This curve, which is drawn on semilog graph paper, readily indicates the medium persistence rating of this phosphor. From the curve, we notice that the relative brightness has decreased to roughly 10 per cent of its maximum value in 20 milliseconds and to about one per cent in 70 milliseconds.



Courtesy RCA

Fig. 7-21.—Luminous intensity characteristic of P1 phosphor as a function of the accelerating potential.

The graph of Fig. 7-21 illustrates the luminous intensity output in candlepower per square centimeter of phosphor surface area with respect to the accelerating potential (which in this case was the second-anode voltage). Each of the four curves shown on this graph was plotted for a different average value of second-anode current per square centimeter of phosphor area. It is seen that for a constant value of second-anode current, the candlepower output per unit area increases with an increase in second-anode voltage. For example, using the 5-μa curve we find that, for second-anode voltage of 6 kilovolts, the luminous intensity is

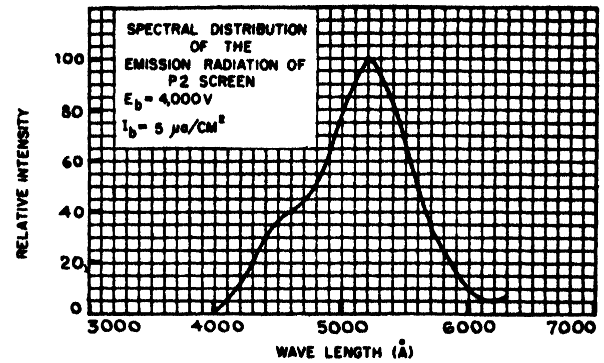


Courtesy RCA

Fig. 7-22.—Luminous intensity characteristic of P1 phosphor as a function of the second-anode current per square centimeter.

0.045 candlepower per square centimeter and, at 13 kilovolts, the luminous intensity is about 0.163 candlepower per square centimeter.

With the second-anode current per unit area plotted against the candlepower per unit area for the same phosphor and with the second-anode voltage as the parameter, a different set of curves results. A typical set of these curves is shown in Fig. 7-22. Examination of these curves reveals that, for a given second-anode voltage, the luminous intensity output increases with an increase in second-anode current.

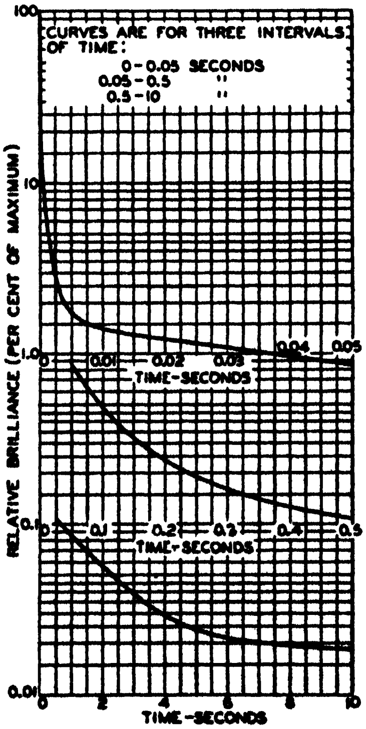


Courtesy Du Mont Labs.

Fig 7-23—Typical spectral characteristic for P2 phosphor.

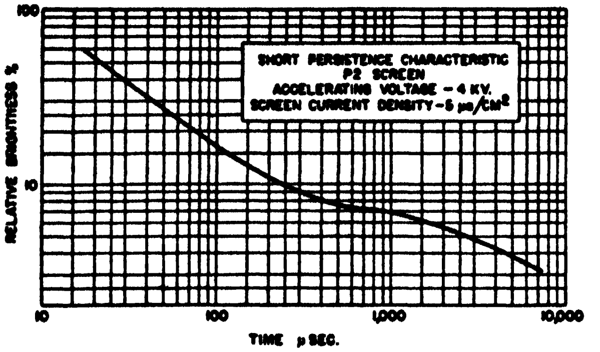
Phosphor P2

The P2 phosphor has a blue-green fluorescent color and a green or yellow-green phosphorescent color with a long persistence Because of its long persistence, this phosphor is chiefly suitable for the observation of transient phenomena having medium writing rates. A typical spectral characteristic for this phosphor is shown in Fig. 7-23. The conditions under which this curve was measured are indicated on the drawing. The



Courtesy Proc. I.R.E.

Fig. 7-24.—Persistence characteristics of typical P2 phosphor for three different time intervals.



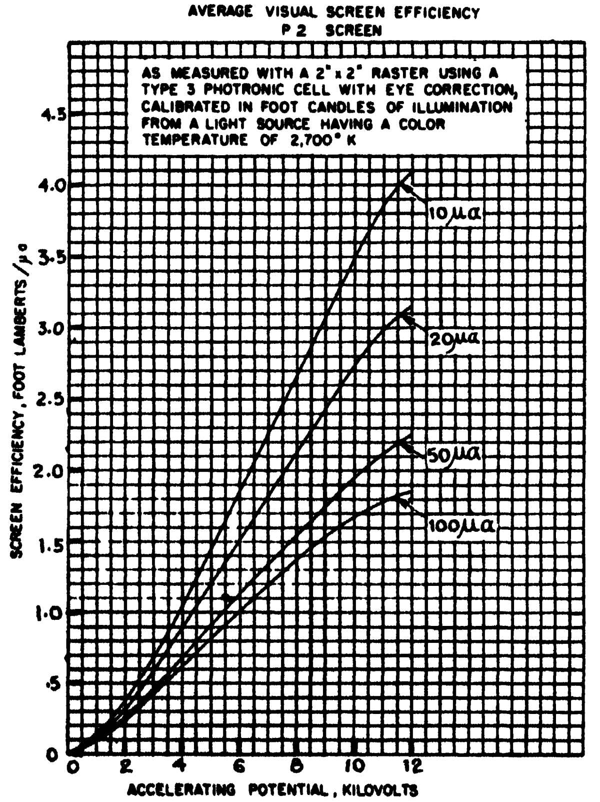
Courtesy Du Mont Labs.

Fig. 7-25.—Short-persistence characteristic of P2 phosphor.

maximum relative intensity occurs at a wavelength of about 5,200 Å, which is approximately in the center of the green region of the visible spectrum. The relative intensity at the lower wavelengths is sufficiently high, due to the broad slope and slight hump of the characteristic, to impart also a blue fluorescent color. At 4,600 Å the relative intensity is still 40 per cent of maximum, and this wavelength falls approximately in the middle of the blue region of the spectrum. This explains the blue-green fluorescent color of this phosphor.

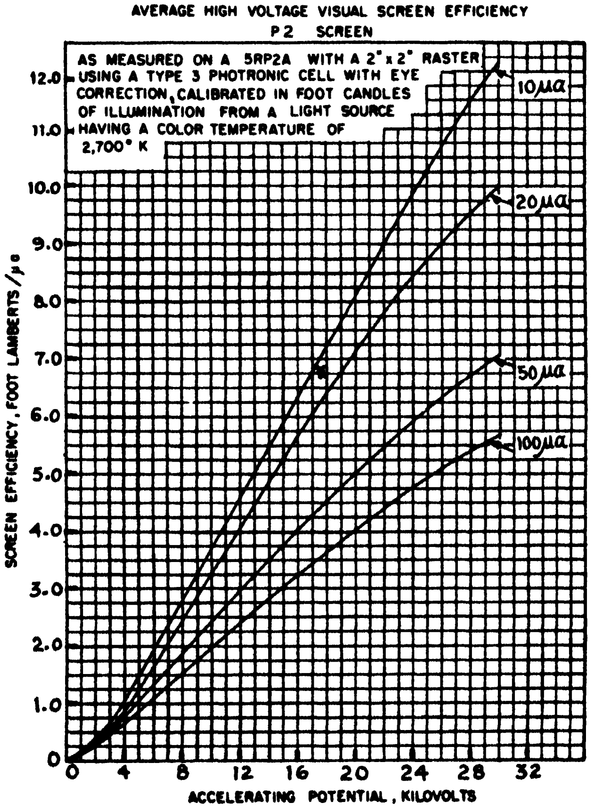
Some of the blue-green color also persists during the time after the excitation voltage is removed, but only for a short period. If a suitable filter is placed across the face of the tube to filter out the long-persistence phosphorescent color (yellow-green), the only afterglow that will be present is the short-persistence blue-green color. Under such conditions, this screen can also be used for general visual observation and photography where a short-persistence screen is required.

In Fig. 7-24 is a curve illustrating a persistence characteristic of a typical P2 phosphor. Because of the long persistence of this phosphor, the curve is shown with the time scale (X-axis) divided into three sections. From this curve, we see that after about 0.05 seconds (50 milliseconds) the relative brilliance has decreased to approximately one per cent of maximum and after ten seconds the relative brilliance is decreased to about 0.02 per cent of maximum. In Fig. 7-25 is a persistence characteristic of another typical P2 screen, but this characteristic is only for the early part of the



Courtesy Du Mont Labs.

Fig. 7-26.—Average visual screen efficiency of a P2 phosphor as a function of the accelerating potential.



Courtesy Du Mont Labs.

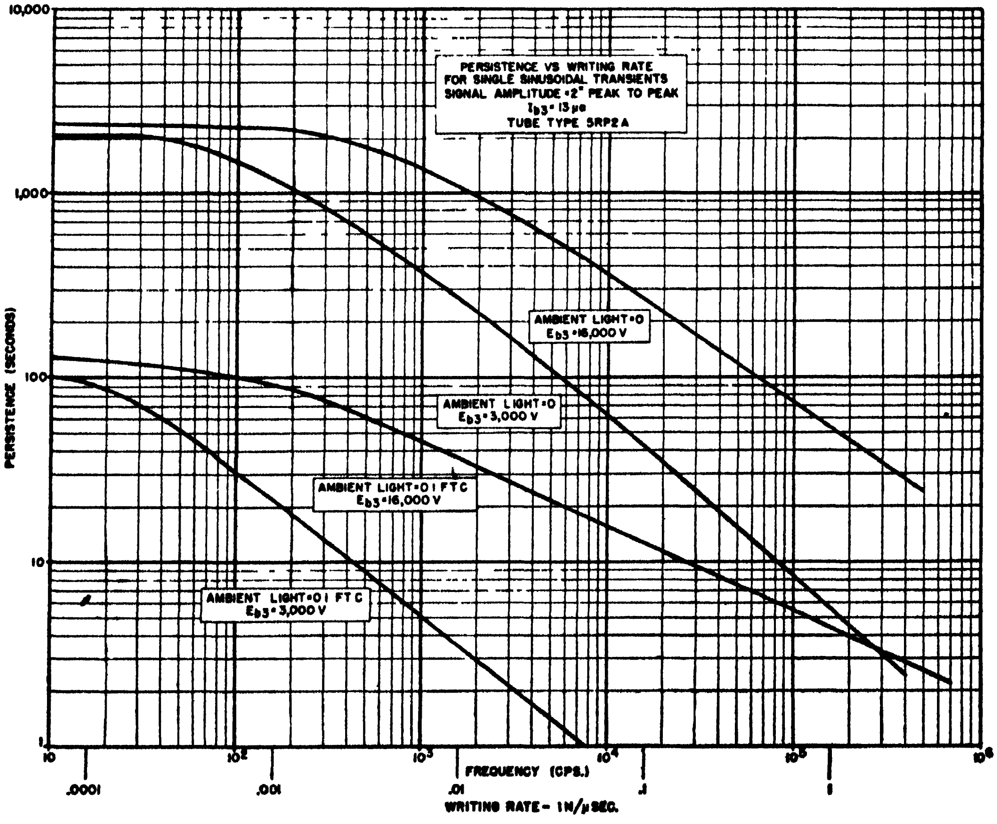
Fig. 7-27.—Average visual screen efficiency of P2 phosphor for high accelerating potentials.

decay. From this chart, we see that after about 8,000 microseconds (8 milliseconds) the brightness has decreased to about 3 per cent of its maximum value.

The *visual* screen efficiency of a typical P2 phosphor is indicated in Figs. 7-26 and 7-27. The graph of Fig. 7-26 is for accelerating potentials of 12 kilovolts and less, while that for Fig. 7-27 includes accelerating potentials to 30 kilovolts. Each curve in these drawings is plotted for a fixed value of beam current. From these curves, it is apparent that there is a great drop in efficiency with increase in screen current.

For long-persistence screens, it is generally useful to include a graph of writing-speed characteristics. The persistence versus writing speed of a typical P2 phosphor is indicated in Fig. 7-28. All curves in this graph were plotted under the input condition of single sinusoidal transients that had a 2-inch peak-to-peak amplitude. The cathode-ray tube used to plot this graph was a Du Mont type 5RP2A and the measured accelerating anode current (i.e., anode number 3, the intensifier electrode) was 13 microamperes. The abscissa or X-axis is calibrated in frequency (cycles per second) as well as writing speeds (inches per microsecond). The frequency and writing-rate scales line up as shown in the graph only for the case of a 2-inch peak-to-peak amplitude. For this signal amplitude, the persistence and writing speeds can be determined directly from the graph for any frequency. For other values of amplitude, the nomograph of Fig. 7-15 would have to be used to determine the writing speed at any one frequency. Once this writing speed is known, the persistence can be taken from the graph, all other things being the same.

You will notice that four curves are shown in Fig. 7-28, each for a different combination of accelerating potential in kilovolts (third-anode voltage) and *ambient light* conditions in foot-candles. By ambient light, we mean the light level surrounding the cathode-ray tube. From these curves we see that for an increase in ambient light (in this case from no ambient light or complete darkness to 0.1 foot-candles of light) the persistence time for any given writing speed decreases. It is also seen that for a given ambient light level, the persistence time increases with increase in accelerating potential.



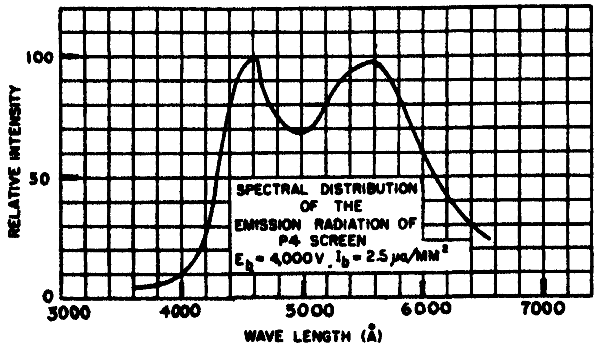
Courtesy Du Mont Labs.

Fig. 7-28.—Persistence versus writing-rate characteristics of a P2 phosphor for different ambient light conditions and accelerating potentials.

At a writing speed of 0.01 inch per microsecond, at no ambient light, the persistence of this phosphor at 16,000 volts accelerating potential is about 1,000 seconds, whereas at 3,000 volts it is about 250 seconds, a difference of 750 seconds, or a 75 per cent reduction in persistence. With some ambient light, the percentage decrease for the same accelerating potentials is even more pronounced. At an ambient light of 0.1 foot-candles and a writing speed of 0.01 inches per second, the persistence for a 16,000-volt accelerating potential will be about 37 seconds, whereas for 3,000 volts the persistence time is about 3.5 seconds, a difference of 33.5 seconds or about a 90 per cent reduction in time. The graph can also be used the other way, that is, for a desired persistence, the required writing speed under different conditions of accelerating potential and ambient light can be determined. Whatever the case may be, it is evident from these graphs that, for increasing writing speed, the persistence decreases.

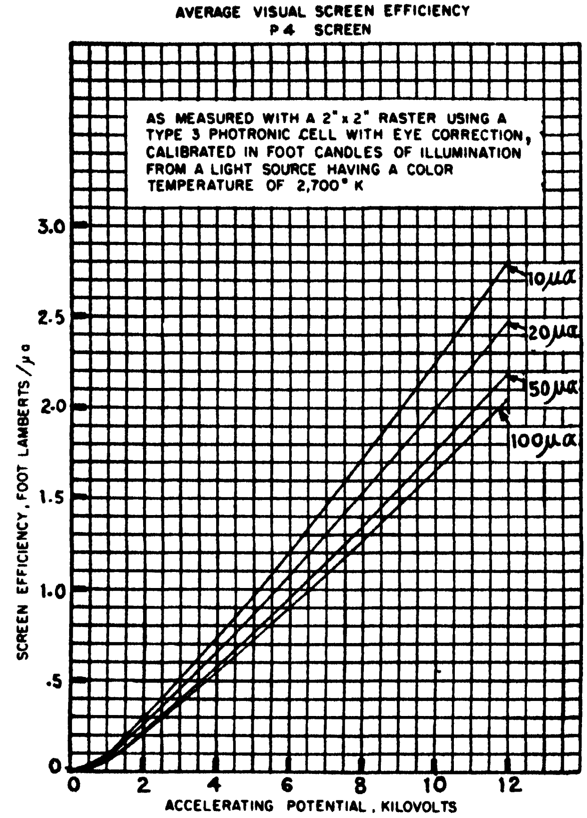
Phosphor P4

The P4 phosphor is employed primarily in television picture tubes. It has a white fluorescence and phosphorescence with a medium persistence. White light is not an individual color but is a mixture of various colors in the visible spectrum. This is evident from a typical spectral characteristic of a P4 phosphor, as shown in Fig. 7-29. In order to cover most of the spectrum, the curve has to be made quite broad. To accomplish this, more than one phosphor is usually employed as the screen material, each having a different spectral characteristic. The combination of these individual characteristics produces the desired curve, similar to that of Fig. 7-29. From this curve it is seen that there is enough relative energy from wavelengths as low as 4,200 Å to wavelengths as high as 6,600 Å to cover all the colors of the spectrum.



Courtesy Du Mont Labs.

Fig. 7-29.—Spectral characteristic of a P4 phosphor, showing two peaks within the visible spectrum.

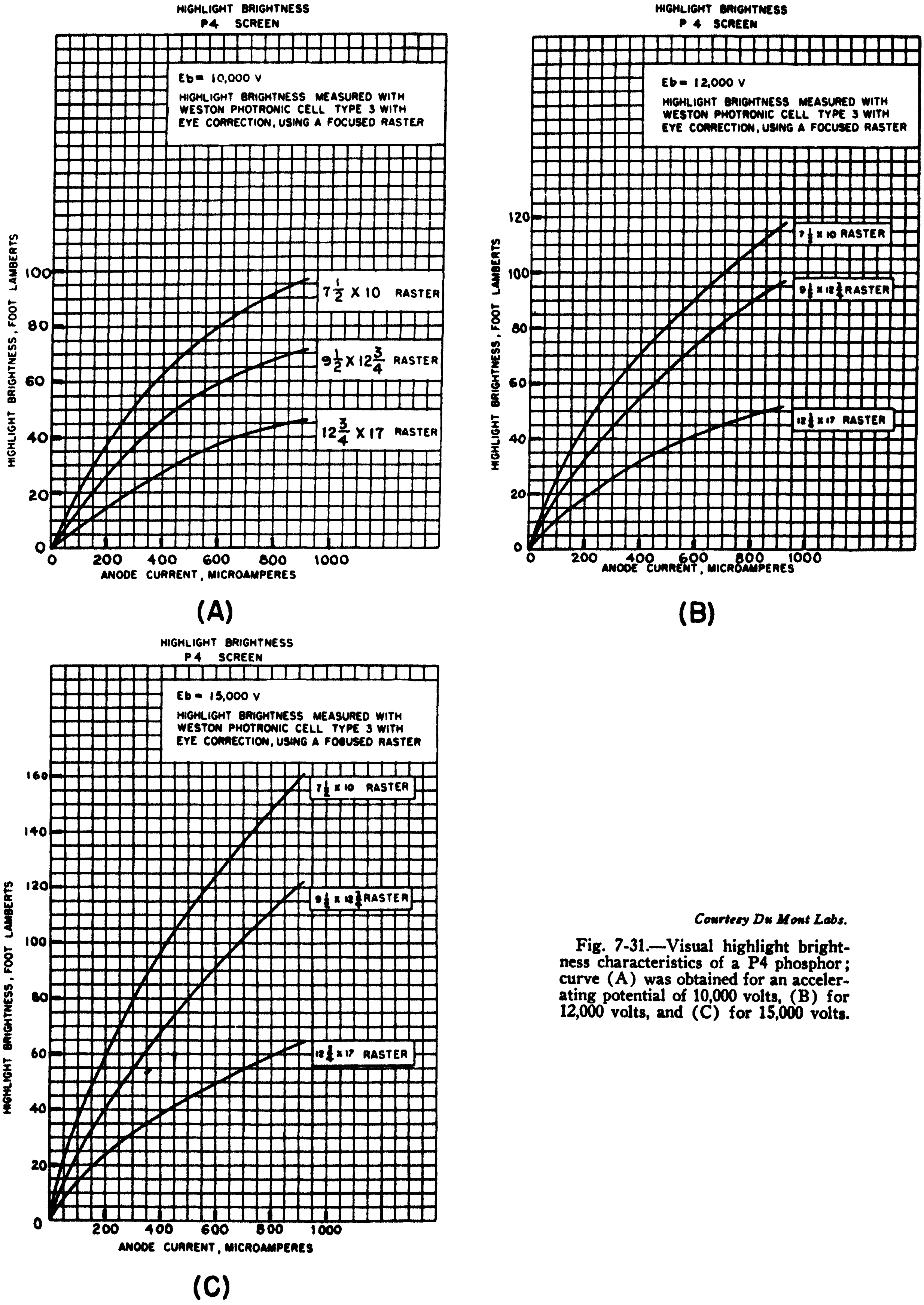


Courtesy Du Mont Labs.

Fig. 7-30.—Average visual screen efficiency of a P4 phosphor as a function of accelerating potential.

The average visual screen efficiency characteristic of a typical P4 screen for a number of different beam currents is illustrated in Fig. 7-30. This graph is for accelerating potentials of 12 kilovolts and less. The brightness is, of course, greater for higher beam currents, but again the efficiency decreases with increasing beam current.

The visual highlight brightness characteristics of a typical P4 phosphor for different values of accelerating potential and for three varying raster sizes are indicated in Fig. 7-31A, B, and C. The different accelerating potentials used are 10,000, 12,000, and 15,000 volts, respectively, as indicated on the graphs. The dimensions (in inches) of the three different size rasters used are also indicated on the graphs; they may be taken to represent different screen sizes of television receivers. From these graphs, we immediately see that, as the accelerating potential increases, the brightness also increases for any one curve (i.e., at any raster size) and any given value of anode current. It is also evident that as the size of the raster increases, the highlight brightness for any one value of anode current decreases.



Courtesy Du Mont Labs.

Fig. 7-31.—Visual highlight brightness characteristics of a P4 phosphor; curve (A) was obtained for an accelerating potential of 10,000 volts, (B) for 12,000 volts, and (C) for 15,000 volts.

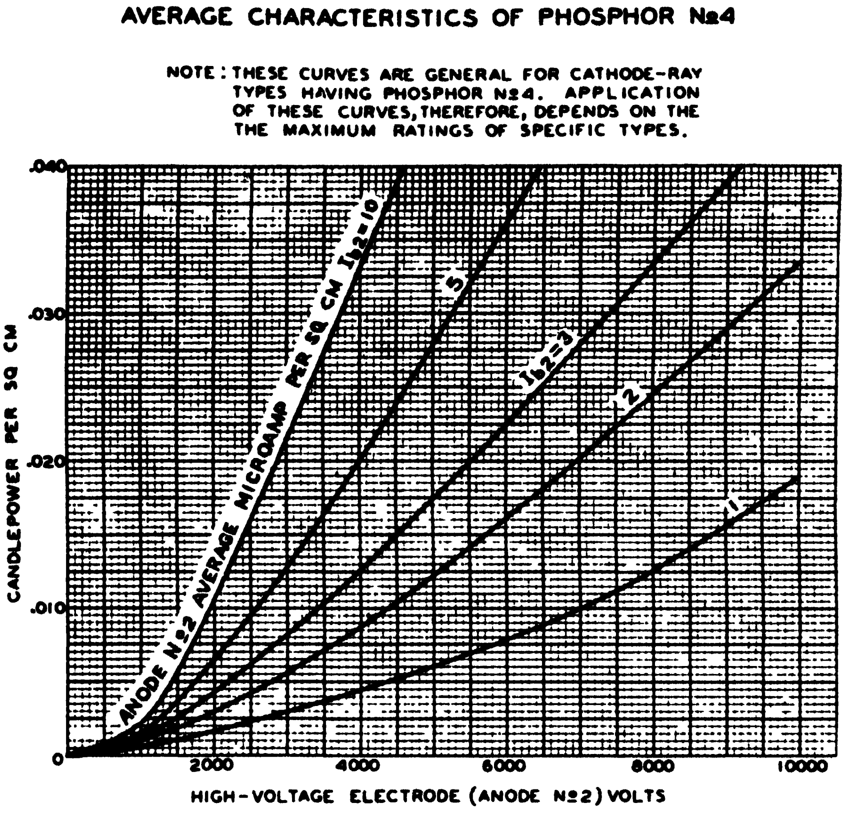


Fig. 7-32.—Average luminous intensity of a typical P4 phosphor as a function of the accelerating potential.

Courtesy RCA

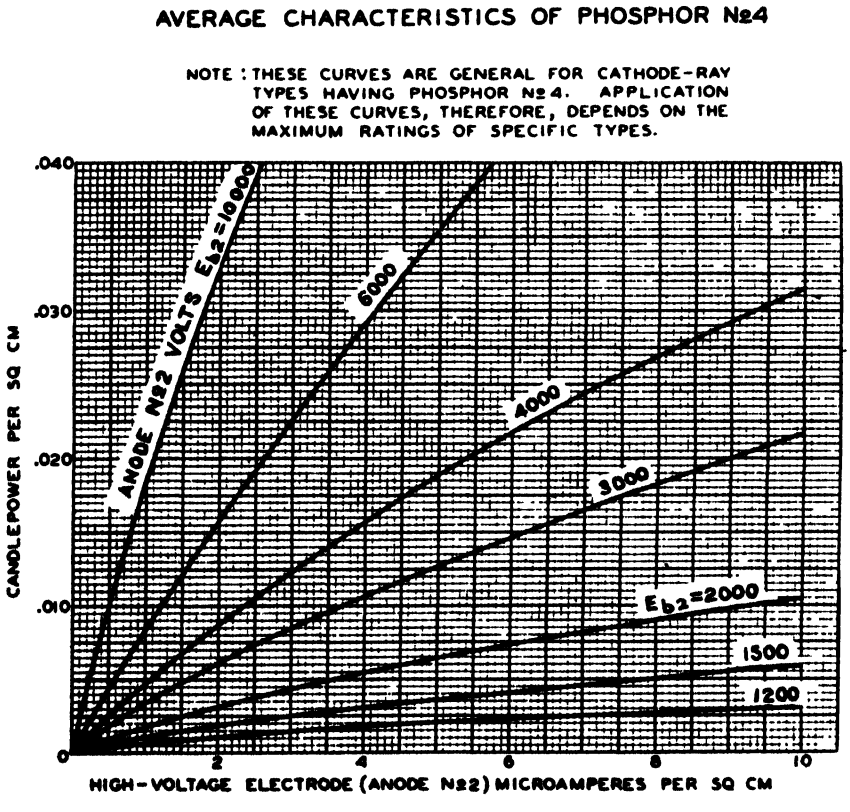
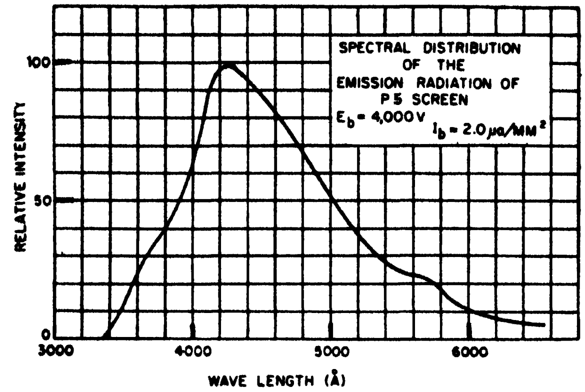


Fig. 7-33.—Average luminous intensity of a typical P4 phosphor as a function of the second-anode current.

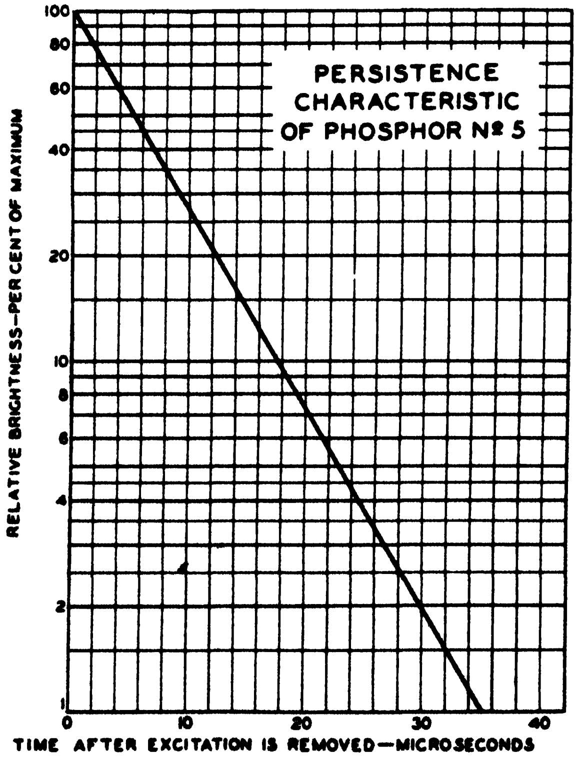
Courtesy RCA



Courtesy Du Mont Labs.

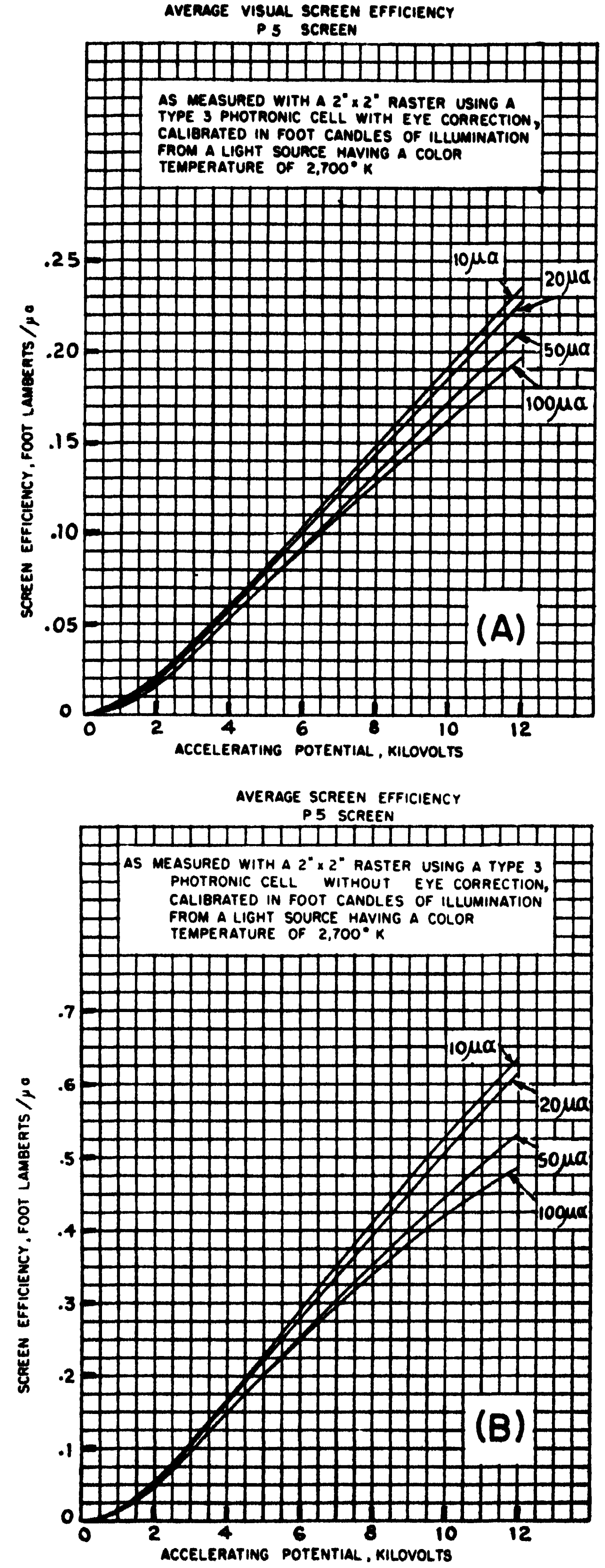
Fig. 7-34.—Spectral characteristic of a typical P5 phosphor.

Average characteristics of a typical P4 phosphor appear in Figs. 7-32 and 7-33. The curves of Fig. 7-32 show luminous intensity in candlepower per square centimeter versus the accelerating potential (in this case the second-anode voltage) for different values of second-anode current per unit screen area. The abscissa of the graph in Fig. 7-33 is the second-anode current per square centimeter and the changing parameter is the second-anode voltage.



Courtesy RCA

Fig. 7-35.—Persistence characteristic of a P5 phosphor, showing the decay curve to 1 per cent of maximum brightness.



Courtesy Du Mont Labs.

Fig. 7-36.—The average visual screen efficiency corrected for eye response is shown in (A) for a P5 phosphor, while the screen efficiency not corrected for the eye's response of the same phosphor is shown in (B).

Phosphor P5

The P5 phosphor has a blue fluorescence and a very short persistence. This latter feature makes this phosphor suitable for photographic recording even when the film is moving at very high speeds. The spectral characteristic of a typical P5 phosphor is shown in Fig. 7-34. Since the blue part of the visible spectrum extends from about 4,300 Å to 4,900 Å, it is readily evident from this spectral characteristic of the P5 phosphor that blue will be the prominent fluorescent color.

The very short persistence of this phosphor is best characterized by the graph of Fig. 7-35. From this graph, we see that for the particular P5 phosphor for which it was plotted, after about 17.5 microseconds, the relative brightness has already decreased to 10 per cent of its maximum value, and in double that time, that is, in about 35 microseconds, the brightness will have decreased to one per cent.

In Fig. 7-36 appear two screen efficiency curves, plotted for accelerating potentials under 12 kilovolts. That at (A) gives the average visual screen efficiency, measured with a photoelectric meter which is corrected for the eye, while part (B) does not have any eye correction. From these two graphs, we can readily see that the comparative screen efficiency with eye correction is much less than that without eye correction. For example, for the 10-μa curve at a potential of 12 kilovolts, the visual screen efficiency, from graph (A), is 0.235 foot-lamberts/μa, whereas for the same curve without eye correction, from graph (B), the efficiency is 0.635 foot-lamberts/μa. In terms of percentage, there is less of a drop in screen efficiency at the higher accelerating potentials, as the beam current is decreased, for the graph with eye correction than there is for the graph without eye correction. For example, for the 10-μa and 100-μa curves, at 12 kilovolts, the visual screen efficiency of graph (A) decreases from 0.235 to 0.197 foot-lamberts/μa, or approximately 16 per cent, whereas for graph (B) the screen efficiency decreases from 0.635 to 0.485 foot-lamberts/μa, or about 31 per cent.

Fig. 7-37 gives the average characteristic of luminous intensity in millicandlepower per square centimeter versus the second-anode voltage (accelerating potential) of a typical P5 phosphor. Each curve is for a specific value of second-anode current per unit screen area as indicated. Note that the luminous intensity for this type of phosphor is lower than that of the previous phosphors for the same voltage and current ratings. The curves of Fig. 7-38 are average characteristics of luminous intensity versus second-anode current per square centimeter of screen area. Each curve in this graph is based on a different second-anode voltage.

Phosphor P7

The P7 phosphor is a cascaded screen of very long persistence. Its uses are similar to that of the P2 screen except that the persistence of the P7 screen is somewhat longer. The fluorescent and phosphorescent color of the P7 screen may vary somewhat according to what is the exact chemical makeup as produced by the different manufacturers. Most P7 screens have a bluish

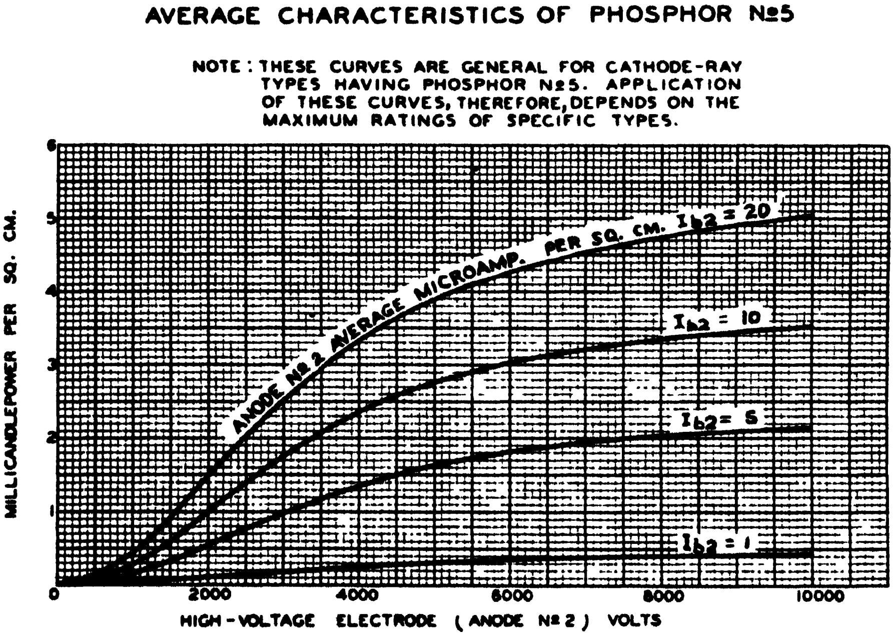
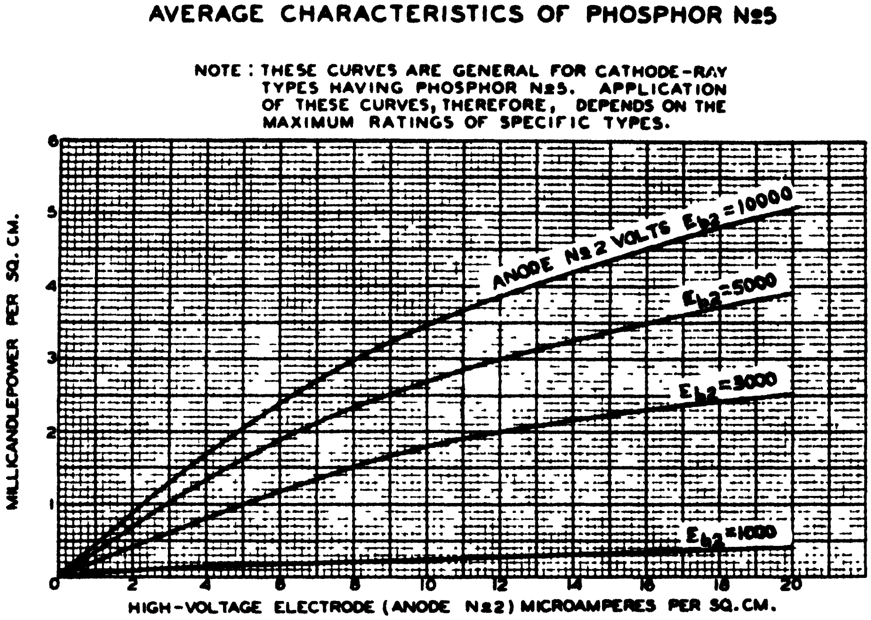


Fig. 7-37.—Average luminous intensity characteristics for a P5 phosphor as a function of the accelerating potential.

Courtesy RCA

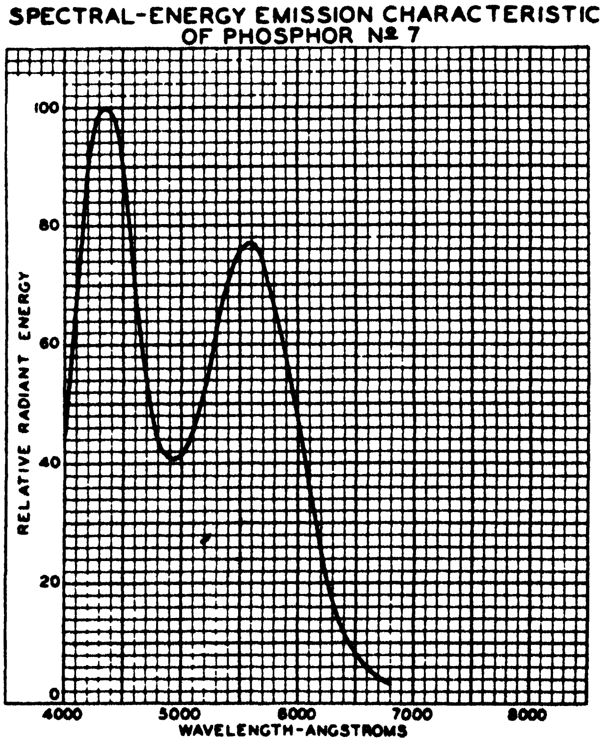


Courtesy RCA

Fig. 7-38.—Average luminous intensity characteristics for a P5 phosphor as a function of the second-anode current.

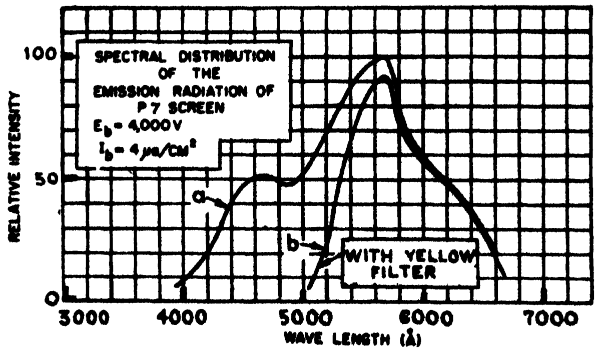
fluorescent color and a yellowish or greenish-yellow phosphorescent color. After excitation, the bluish color persists for only a short interval of time.

To indicate how spectral characteristics of this phosphor may vary with different manufacturers, the two different P7 spectral characteristics are included here. Fig. 7-39 shows the spectral characteristic for the P7 phosphor manufactured by RCA. Note that two peaks exist in this curve, but that maximum sensitivity occurs at about 4,300 Å, in the blue region of the spectrum, while the other peak, which is about 77 per cent of maximum energy. occurs at a wavelength of about 5,600 Å in the yellow region of the spectrum. Curve *a* in Fig. 7-40 is the spectral characteristic of the P7 screen manufactured by Du Mont. The two peaks occur in approximately the same region of the spectrum as in Fig. 7-39, but for curve *a* of Fig. 7-40, the maximum intensity occurs at the second peak (yellow part of the spectrum), rather than at the first peak. The resultant fluorescence as stated by each manufacturer is, however, the same bluish color.



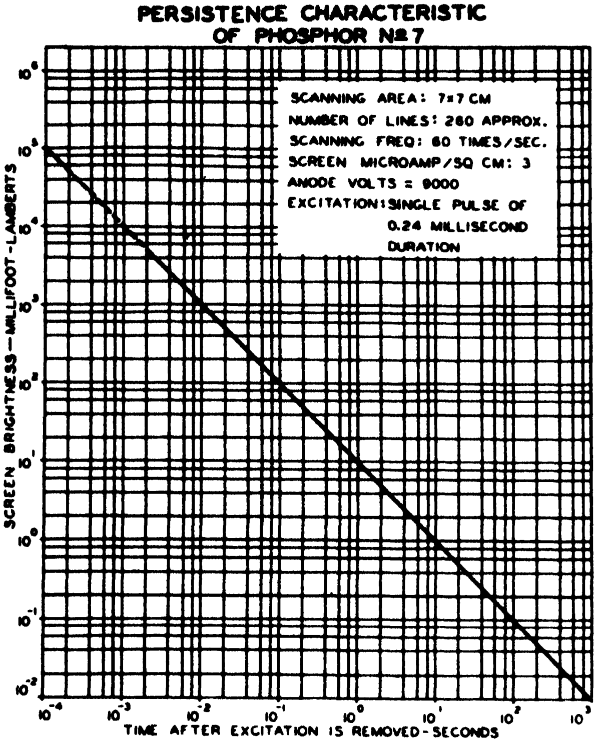
Courtesy RCA

Fig. 7-39.—Spectral energy characteristic for a typical P7 phosphor, showing two peaks, with the maximum sensitivity in the yellow region.



Courtesy Du Mont Labs.

Fig. 7-40.—Spectral energy characteristic of the Du Mont P7 phosphor with and without yellow filter; compare this characteristic with the one shown in Fig. 7-39 for an RCA type P7 phosphor.



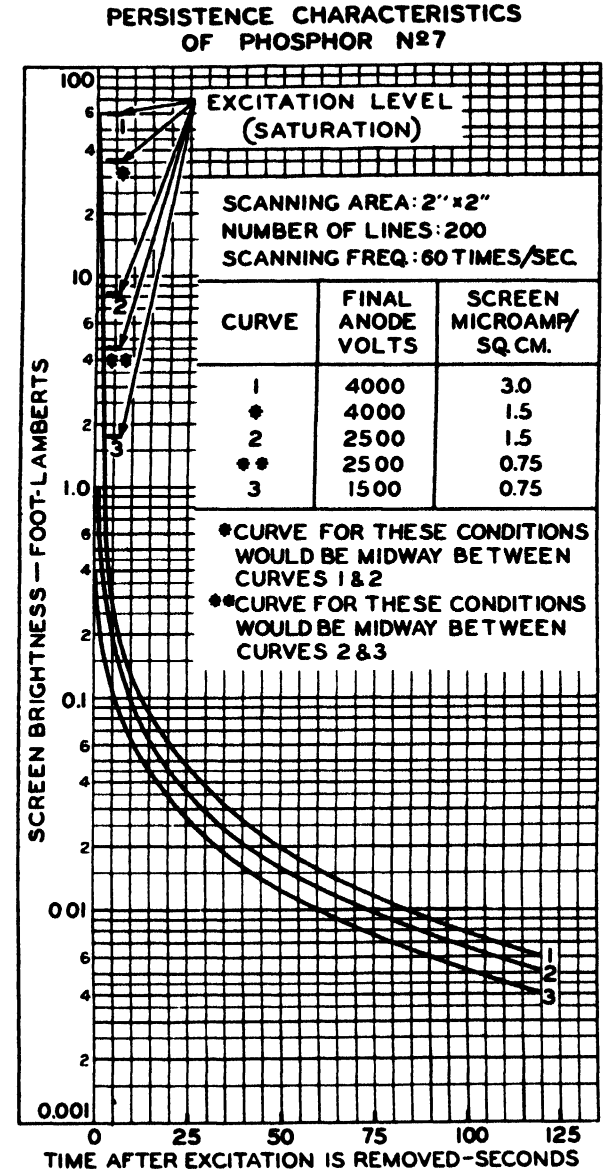
Courtesy RCA

Fig. 7-41.—Typical persistence characteristic of a P7 phosphor.

Curve *b* of Fig. 7-40 is the resulting spectral characteristic of the Du Mont P7 phosphor when a yellow filter is used. Note that by use of this filter the light of lower wavelengths, 5,000 Å and below, is cut out. The reverse effect can be had if a blue filter is used instead of the yellow one.

An average persistence characteristic of a typical P7 phosphor appears in Fig. 7-41. It is log-log plot (that is, both scales are logarithmic), with the ordinate or *Y*-axis calibrated in brightness values of millifoot-lamberts. The conditions under which this graph was plotted are indicated on the drawing. The relative brightness for the later stages of decay time of this phosphor, for somewhat different conditions of operation than those indicated in Fig. 7-41, is plotted in Fig. 7-42. Although only three individual curves are shown in this latter graph, note that two other curves for other stipulated conditions, would fall midway between curves 1 and 2, and 2 and 3, respectively.

The writing-rate characteristic for the Du Mont P7 phosphor for the same accelerating voltage and ambient light conditions, as previously shown in Fig. 7-28 for the P2 screen, appears in Fig. 7-43, with the exception, of course, that a 5RP7A tube is used. Both the graph of Fig. 7-28 and that of Fig. 7-43 are plotted to the same scales.



CURVE	FINAL ANODE VOLTS	SCREEN MICROAMP/ SQ. CM.
1	4000	3.0
*	4000	1.5
2	2500	1.5
**	2500	0.75
3	1500	0.75

*CURVE FOR THESE CONDITIONS WOULD BE MIDWAY BETWEEN CURVES 1 & 2

**CURVE FOR THESE CONDITIONS WOULD BE MIDWAY BETWEEN CURVES 2 & 3

Courtesy RCA

Fig. 7-42.—Typical persistence characteristics of a P7 phosphor under various conditions of final anode voltage.

Phosphor P11

The P11 phosphor is similar to the P5 phosphor, having a blue fluorescence and a short persistence. It is also useful for photographic work, but, having a somewhat higher persistence than the P5 phosphor, it cannot be used at very high film speeds. The P11 phosphor has a high brightness characteristic. A typical spectral characteristic of a P11 phosphor appears in Fig. 7-44 for a fixed accelerating voltage and beam current per unit area. It is seen that the highest intensity occurs at wavelengths well within the blue region of the visible spectrum.

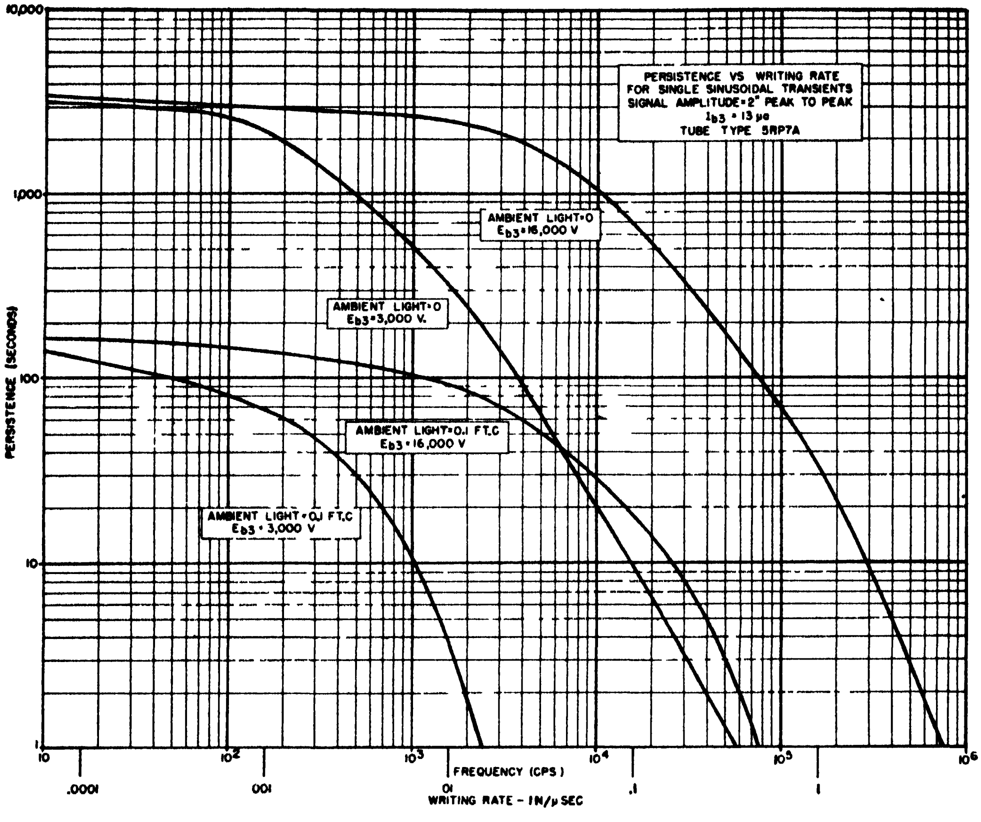
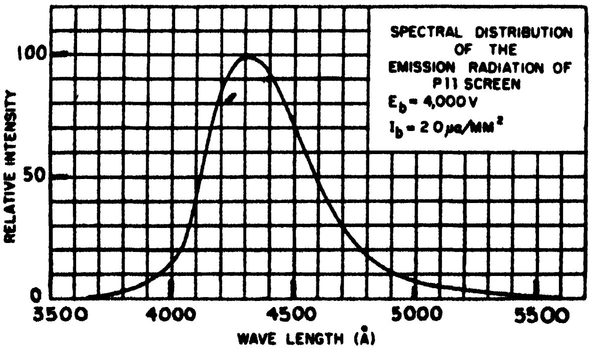
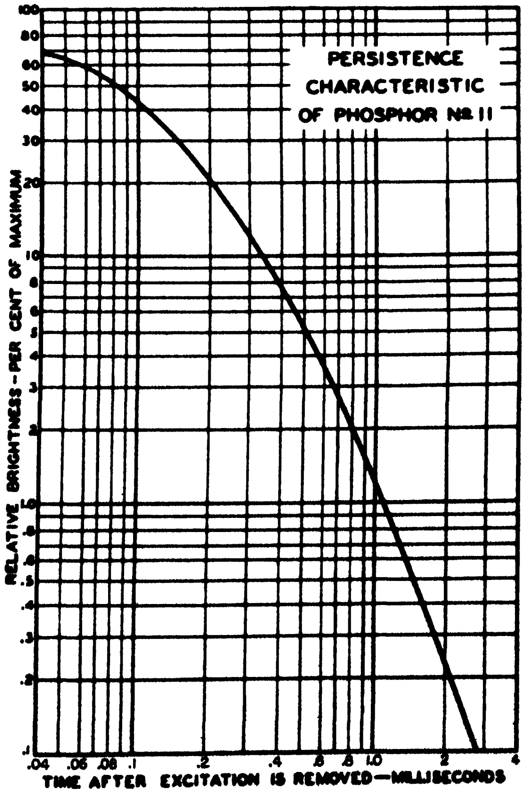


Fig. 7-43.—Writing-rate characteristics for the Du Mont P7 phosphor, using a 5RP7A tube. *Courtesy Du Mont Labs.*

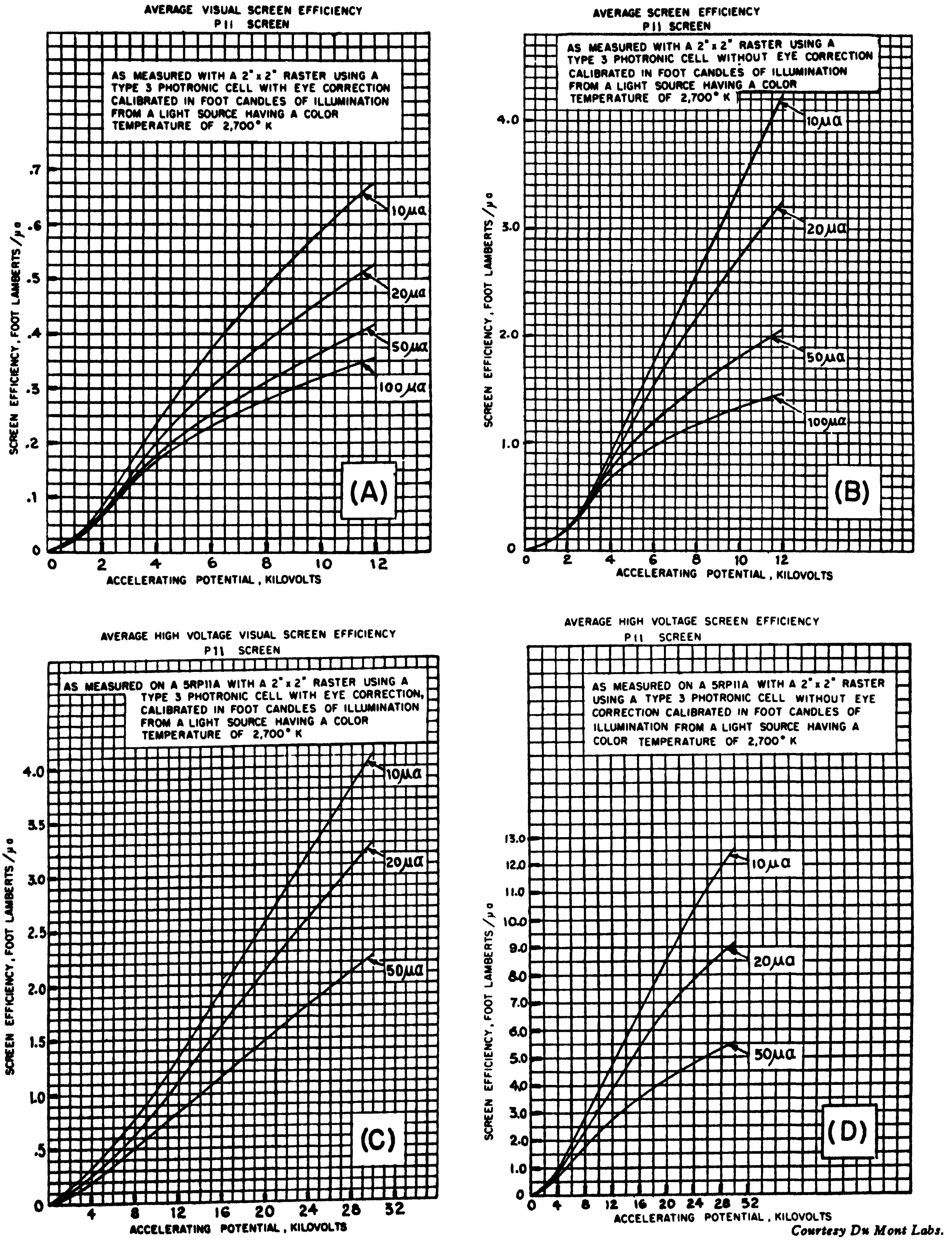
A typical persistence characteristic of the P11 phosphor appears in Fig. 7-45. This characteristic is a log-log plot of relative brightness versus time in milliseconds. Comparison between this persistence characteristic and that for phosphor P5 (see Fig. 7-35) will readily show that the P5 phosphor has a considerably shorter persistence than the P11 phosphor. From Fig. 7-35, we see that after about 40 microseconds, the relative brightness will have decreased to less than one per cent of the maximum brightness, whereas for phosphor P11, Fig. 7-45, at 40 microseconds (that is, at 0.04 millisecond), the relative brightness has



Courtesy Du Mont Labs.
Fig. 7-44.—Typical spectral characteristic of a P11 phosphor.



Courtesy RCA
Fig. 7-45.—Typical persistence characteristic of a P11 phosphor.



Courtesy Du Mont Labs.

Fig. 7-46.—Curves representing average screen efficiencies of a P11 phosphor; (A) and (B) are for low accelerating voltages, with (A) taking into account eye correction; (C) and (D) are for high acceleration voltages, with (C) taking into account eye correction.

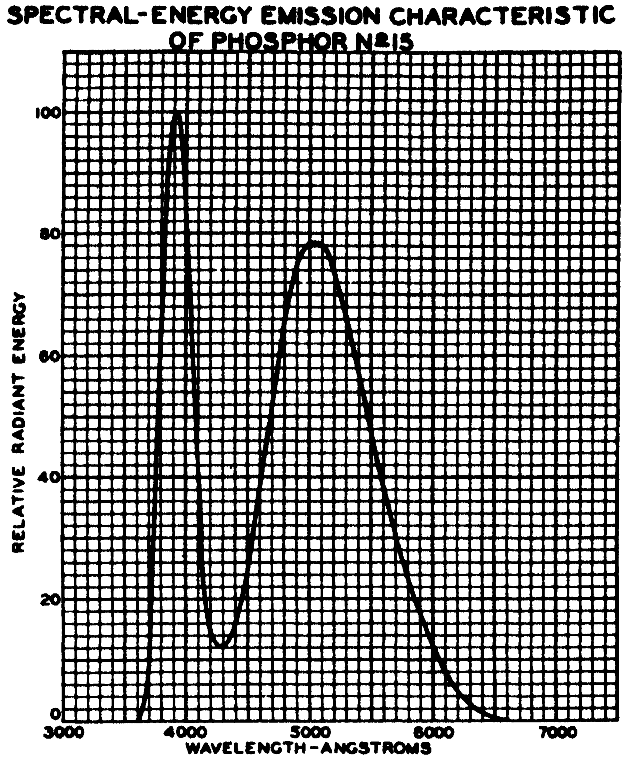
only decreased to about 70 per cent of its maximum value. For phosphor P11, it will take about 1.1 milliseconds (1,100 microseconds) before the relative brightness will have decreased to one per cent of maximum.

A series of average screen-efficiency characteristics for a typical P11 phosphor is illustrated in Fig. 7-46. Graphs (A) and (B) are for low accelerating voltages (below 12 kilovolts) and graphs (C) and (D) are for higher accelerating voltages, to about 30 kilovolts. From the conditions under which each graph has been plotted, you will note that graphs (A) and (C) are for average *visual* screen efficiencies, that is, they take into account eye correction, while graphs (B) and (D) do not correct for the response of the human eye. As in the previous average screen-efficiency curves, the visual efficiency is much less than the uncorrected efficiency.

Phosphor P15

The P15 phosphor is a relatively new screen material as compared to the other phosphors. This phosphor has a very short persistence and fluoresces blue-green and near ultraviolet. The latter region, of course, is not visible. The fluorescent characteristics of a typical P15 phosphor appear in Fig. 7-47. Note that there are two peaks in this curve, the higher peak (maximum energy) occurring at a wavelength of about 3,900 Å, and the smaller peak (at 78 per cent of maximum energy) occurring at about 5,000 Å. The trough between the two peaks occurs at 4,300 Å, which is the upper wavelength end of the violet part of the spectrum. The spectral bandwidth about the first peak, for radiant energy about 12 per cent of maximum, extends between 3,700 Å to 4,300 Å, which is within the ultraviolet and violet part of the spectrum. Since wavelengths around 4,000 Å are practically nonvisible, only an infinitesimal amount of visible light is contributed by this part of the spectral characteristic. The remainder of the spectral characteristic falls primarily in the blue and green regions of the spectrum, which is the fluorescent and phosphorescent color of the phosphor.

As mentioned, the over-all persistence of this screen is very short, but the violet and ultraviolet part of the energy has an even shorter persistence than the blue-green portion of the spectrum. This extremely short persistence of the violet and ultraviolet light makes the phosphor suitable for high-speed scanning requirements. In fact, the P15 phosphor is used in flying-spot cathode-ray tubes (such as RCA type 5WP15, explained in Chapter 21), where the violet and ultraviolet radiation is primarily used. By restricting operation to this radiation, blurring or trailing in the reproduced picture is reduced a great deal.



Courtesy RCA

Fig. 7-47.—Spectral characteristic of a P15 phosphor.

Comparison of Phosphors

It is often desirable to compare different screens by plotting the same characteristics of each screen on one graph. Some of the characteristics which are often compared are the persistence, writing rate, and screen efficiency. In this section, we will illustrate a few of these graphs which have been published in different periodicals. For further comparative information about these phosphors, it is suggested that the bibliography be consulted.

Comparison of Persistence Characteristics

The graph of Fig. 7-48 compares the persistence characteristics of P1, P2, and P5 type phosphors which have medium-, long-, and short-persistence ratings, respectively.[6] The graph is a plot of relative brilliance on a logarithmic scale to decay time in seconds on a linear scale. From this graph, curve *C*, which is for

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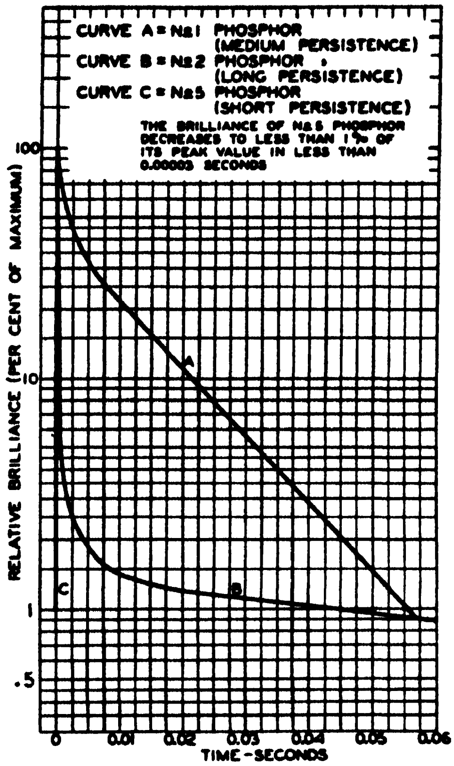
[6] Perkins, T. B., and Kaufman, H. W., "Luminescent materials for cathode-ray tubes," *Proc. I.R.E.*, vol. 23, p. 1329, November, 1935.



Courtesy Proc. I.R.E.

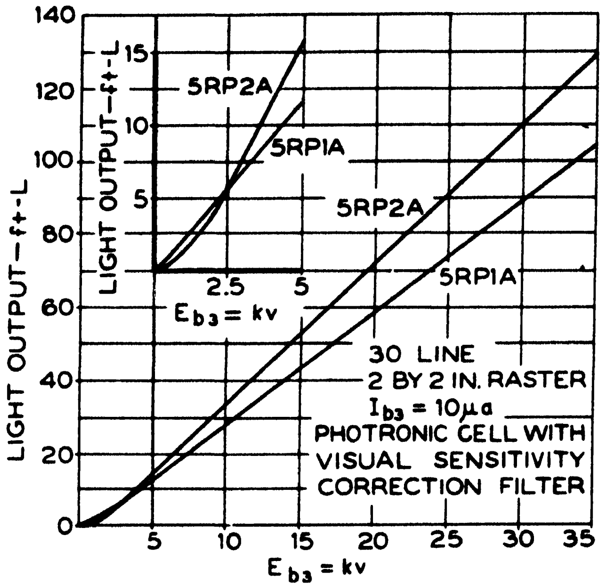
Fig. 7-48.—Comparison of persistence characteristics; curve *A* for P1 phosphor, curve *B* for P2 phosphor, and curve *C* for P5 phosphor.

phosphor P5, appears to decrease to one per cent of maximum brilliance in zero time. Actually this decrease, as noted on the graph, occurs in about 30 microseconds (0.00003 seconds), but it cannot be accurately plotted on the time scale used. The long-persistence characteristic of the P2 screen first becomes evident after about 0.0005 seconds, since the relative brilliance of the curve decreases very slowly after this time.

Comparison of Average Light Output

The graph of Fig. 7-49 compares the light output, calibrated in foot-lamberts, of a P1 and a P2 screen at different values of accelerating potential.[7] Both curves were plotted under the same conditions, that is, a 30-line, 2- by 2-inch raster was used in each type screen; the beam current was 10 μa, and the curves are corrected for visual observation. Du Mont types 5RP1A and 5RP2A cathode-ray tubes were used for the plot of these curves. The cross-over point between the two curves that occurs at low voltages is shown in enlarged form in the upper left-hand corner of the drawing.

Comparison of these two curves at high accelerating potentials tells us that the P2 screen has a greater light output than the P1 screen. For example, at 20 kv, the P2 screen has an output of 72 foot-lamberts, whereas the P1 has an output of 58 foot-lamberts, a difference of 14 foot-lamberts. At 2.5 kv, the light output from both screens is about the same, approximately 5.5 foot-lamberts. Below 2.5 kv, the P1 screen has a greater light output. Remember, however, that the P1 screen has a medium persistence, and that the P2 screen has a long persistence.

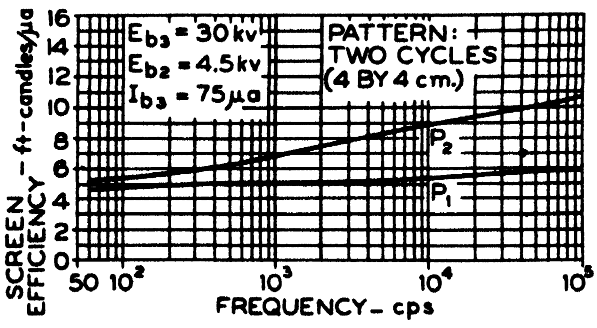


Courtesy Du Mont Labs.

Fig. 7-49.—Comparison of light output of a P1 and P2 screen.

Comparison of Screen Efficiencies

While on the subject of comparing P1 and P2 phosphors, let us compare their screen efficiencies (in foot-lamberts/μa) with respect to the frequency of the applied signal. This graph is illustrated in Fig. 7-50. The intensifier-electrode voltage for the 5RP tubes used was 30 kv and its current drain was 75 μa. The second-anode voltage was 4.5 kv. A two-cycle pattern measuring 4 by 4 cm was used for the measurement. This



Courtesy Du Mont Labs.

Fig. 7-50.—Comparison of screen efficiencies of a P1 and P2 phosphor.

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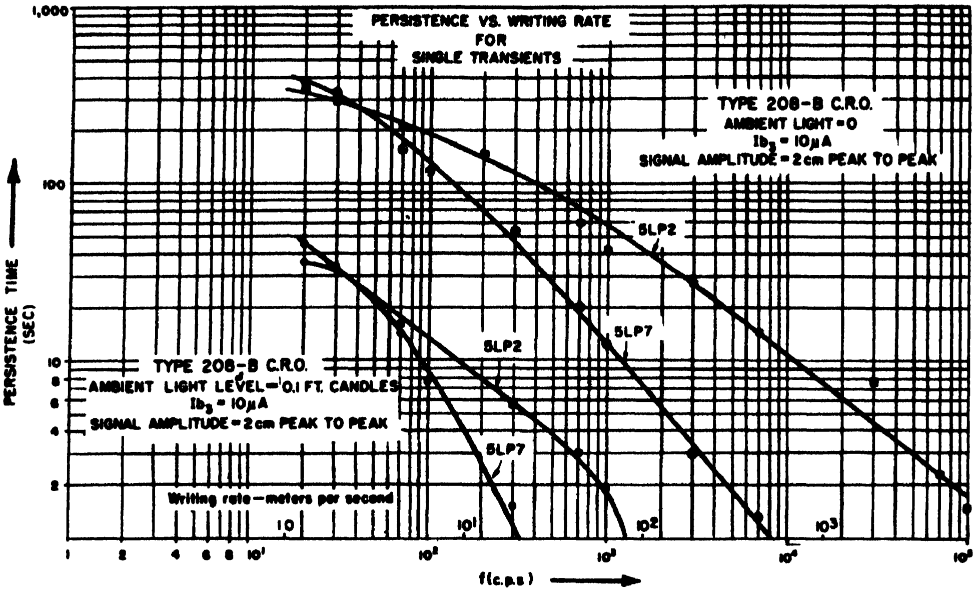
[7] Feldt, R., "An all purpose screen for high-voltage cathode-ray tubes," *The Oscillographer*, vol. 9, p. 3, July-August, 1947.



Courtesy Du Mont Labs.

Fig. 7-51.—Comparison of the writing-rate characteristics of 5LP2 and 5LP7 cathode-ray tubes, beam current of 10 μa.

graph tells us that, for increasing signal frequencies, the screen efficiency of the P2 phosphor will increase at a greater rate than that of the P1 phosphor.

Comparison of Persistence versus Writing-Speed Characteristics

The writing-speed characteristic of long-persistence screens is very useful. Comparison of these characteristics for long-persistence screens is helpful in determining which one is best suited for the study of single transient phenomena. Two graphs will be studied here, both comparing the P2 and P7 long-persistence screens. All curves to be illustrated are plots of actual experiments.

The graph of Fig. 7-51 illustrates the writing-rate characteristics at the low accelerating potential of 1,100 volts.[8] A 5LP2 and a 5LP7 cathode-ray tube were used in a Du Mont type 208-B oscilloscope for the tests. The same signal amplitude (2 cm peak-to-peak) and beam current (10 μa) were used for all curves. The upper two curves were measured in total darkness, whereas the two bottom curves were measured under an ambient light condition of 0.1 foot-candles. For frequencies above about 35 cps, the cross-over point, the P2 phosphor for either light condition, has a longer persistence than the P7 at any one frequency or writing rate. Below 35 cps, or below a writing speed of about 2 meters per second, the P7 phosphor is somewhat superior.

The other graph appears in Fig. 7-52. Eight different curves appear on this graph. Each is measured for a 2-cm peak-to-peak signal amplitude and a beam current of 13 μa. A 5RP2 and a 5RP7 cathode-ray tube were used for this experiment. The four curves starting at the upper part of the persistence scale were measured in total darkness, and the bottom four curves were measured with the ambient light being 0.1 foot-candles. These curves are for higher accelerating potentials than those shown in Fig. 7-51. In each group of four curves, two were plotted with an accelerating potential of 3,000 volts (intensifier electrode), and the other two at a potential of 16,000 volts.

From this latter graph, as well as from the previous one, we find that the ambient light conditions greatly influence the persistence of the phosphors. From Fig. 7-52, it is evident that at higher accelerating potentials, the persistence of the P2 screen is lower than that of the P7 at low frequencies, while it becomes greater than that of the P7 screen at considerably higher frequencies. For example, in total darkness, the P7 screen at 16,000 volts has a longer persistence than the P2 screen for all frequencies up to 100 kc, but at 3,000 volts and the same light conditions, the persistence of the P7 screen is greater only for frequencies up to 2,500 cps. This is immediately evident from the cross-over points of the upper group of curves. For ambient light conditions of 0.1 foot-candles, however, the cross-over points for accelerating potentials of 16,000 and 3,000 volts occur at frequencies of about 26 and 1.7 kc, respectively. This means that the P2 phosphor has a

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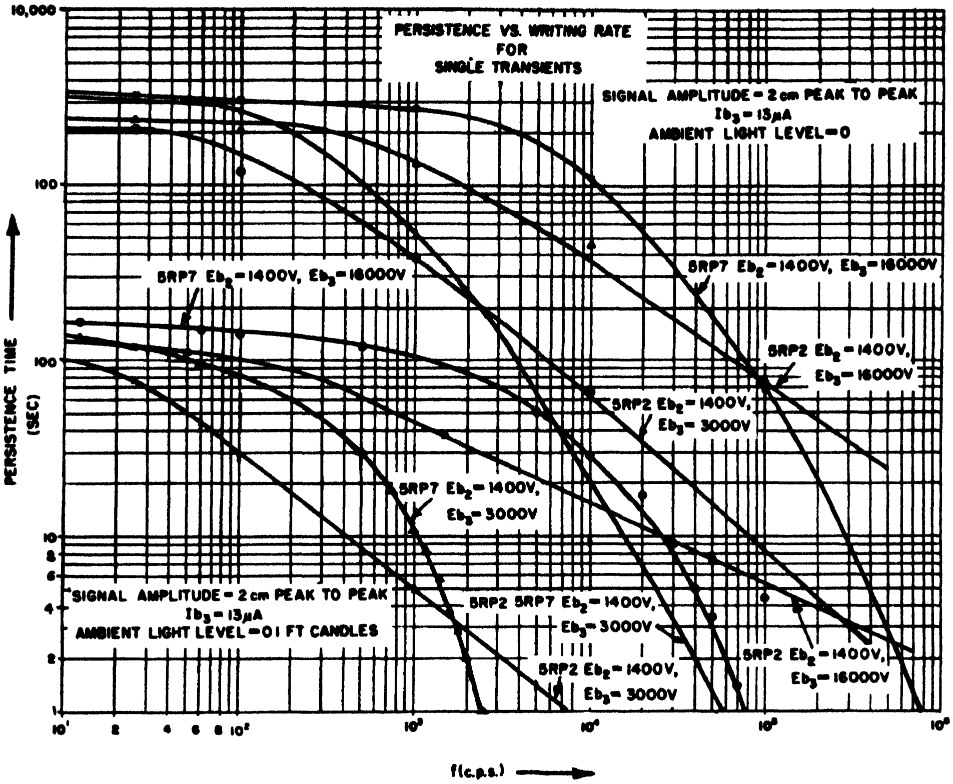
[8] Feldt, R., "Luminescent screens for cathode-ray oscillography," *The Oscillographer*, vol. 11, pp. 8-9, April-June, 1949.



Courtesy Du Mont Labs.

Fig. 7-52.—Comparison of the writing-rate characteristics of 5RP2 and 5RP7 cathode-ray tubes, beam current of 13 μa.

longer persistence than the P7 phosphor at frequencies higher than those for the same conditions of operation.

A general over-all comparison of the P2 and P7 phosphors for the observation of single transients shows that the P2 phosphor has a better persistence at higher frequencies, whereas the P7 phosphor has a better persistence at lower frequencies.

SPECIAL PURPOSE SCREENS

Production of television picture tubes having improved brightness and contrast is one primary aim of cathode-ray-tube manufacturers. To accomplish this and other goals, manufacturers concentrate their efforts on improving the screens of cathode-ray tubes. In this section, we will study some of these special screens, including metal-backed screens, "black" or "dark" faced tubes, and the "skiatron" or "dark-trace" screen phosphor.

Metal-Backed Screens

The following information on metal-backed screens is reprinted from the reference article listed in the footnote.[9] We refer to this article because we believe it contains all the information necessary to explain how and why metal-backed screens help produce better light output.

"One of the outstanding quests in the cathode-ray-tube field has been the search for means of increasing the brightness of the pictures on the face of the tube. Previous methods consisted primarily of efforts to increase beam power—that is, raising the voltage and increasing the current by improvements in electron optics—as well as a search for luminescent materials with greater efficiency in converting beam power into light. The most recent step in increasing light output is the application of a light-reflecting metallic layer on the beam side of the fluorescent screen.

"Many practical tubes with metallic layers on the screens were built and used as far back as 1940. However, these tubes were limited to high-voltage operation and the metallic layers did not possess the light-reflecting properties which characterize the new metal films. The advantages of having a thin reflecting layer have long been anticipated and, to a limited extent, observed in the laboratory. It is only recently, however, that methods have been developed which will make such tubes possible and practical.

"Before showing how this is accomplished, it is worthwhile to review briefly the pertinent part of the

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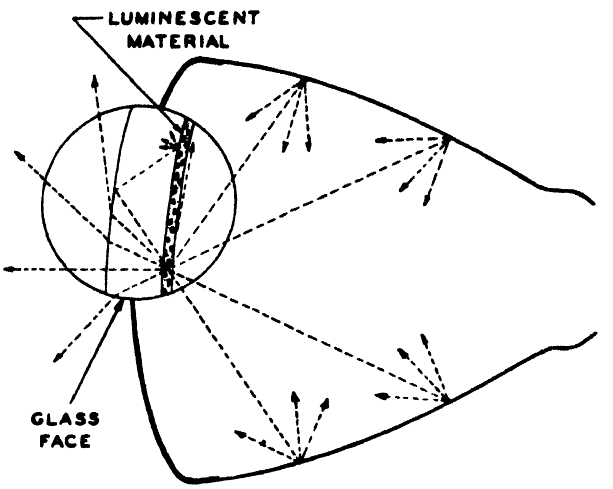
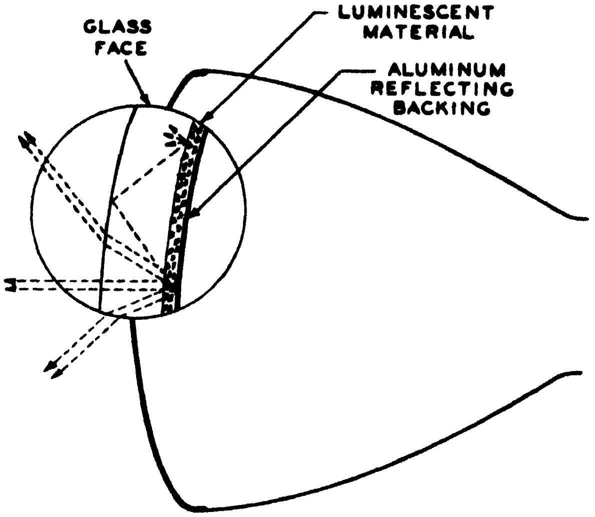
[9] Epstein, D. W., and Pensak, L., "Improved cathode-ray tubes with metal backed luminescent screens," *RCA Rev.*, vol. 7, pp. 5-10, March, 1946.



Courtesy RCA

Fig. 7-53.—Conventional kinescope showing typical distribution of light from a spot.



Courtesy RCA

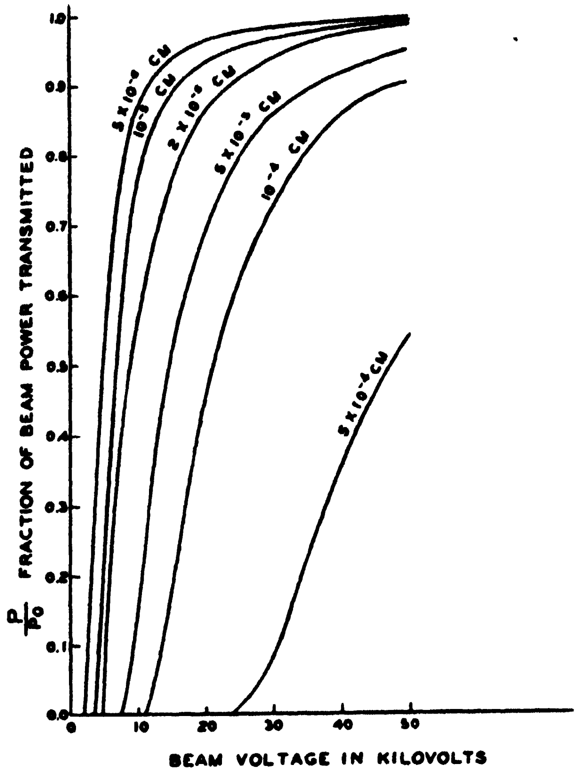
Fig. 7-54.—Kinescope with reflecting metallic film showing gain due to improved distribution of light.

state of affairs at the luminescent screen in a conventional kinescope. This is shown schematically in Fig. 7-53. The region in the circle is a greatly magnified and somewhat distorted diagram of a small section of the tube face of which one element is fluorescing. Generally, at least 50 per cent of the light generated in the screen is emitted towards the electron gun in the tube. Another 15-25 per cent is lost by total internal reflection in the glass of the tube face. Thus only about 25-35 per cent of the total light generated is emitted in the forward direction to constitute the useful light output of the tube. It should also be pointed out here that some of the wasted light is harmful in that it is scattered back onto the screen to set a limit on possible contrast in the picture. There are several mechanisms for this. One is the back-scattering of the light which strikes the inside walls of the tube; although the light is largely absorbed by the blackening on the wall, some of it comes back to the screen. Another is the light from one portion of the screen which can illuminate other regions directly because of the curvature of the face. Some of the totally reflected trapped light in the glass is reflected back onto the screen and is scattered causing what is known as halation.

"Fig. 7-54 shows a tube whose screen is covered with an electron-pervious, but light-reflecting, metallic layer. Now it is seen that the light which previously would go towards the rear (electron gun) is reflected forward into the direction of viewing. Thus without an increase of light generated, the efficiency of conversion of electron beam power into useful light has been increased. At the same time, some of the limitations on contrast have been removed, that is, the back-reflected light and the effects due to curvature of the face. Experiments show that the large area contrast is considerably improved by a factor of three to ten times; the detail contrast, being primarily limited by halation, is only slightly improved.

"The properties which this metallic layer should possess are: (1) it should be thin enough and of the right kind of metal to cause negligible absorption of the electron beam at the desired operating voltages; (2) it should be opaque, relatively smooth, and highly light-reflecting, so as to act as a mirror; (3) it should have sufficient conductivity to conduct the full beam current; (4) it should be strong enough to withstand the stresses due to effect of the focused electron beam; (5) it should be durable enough to be able to withstand the necessary subsequent processing of the tube; and (6) it should be of a metal that will not chemically react with the luminescent screen material.

"The metal chosen to work with is aluminum, because it combines properties which provide the best compromise in meeting the above conditions. It is easily applied by evaporation and does not affect luminescent screens. Its ability to meet condition (1) is indicated in Fig. 7-55 where there are shown a group of calculated curves giving the fraction of electron beam power that is passed by films of various thicknesses as a function of the initial beam voltage. It will be noticed that a 10,000-volt beam will retain only 15 per cent of its incident energy on passing through an aluminum film 5,000 Å thick. A film 2,000 Å thick will pass 57 per cent and a film 1,000 Å will pass about 77 per cent of the energy. If we assume that the effect of the mirror is to double the apparent brightness, then it is evident



Courtesy RCA

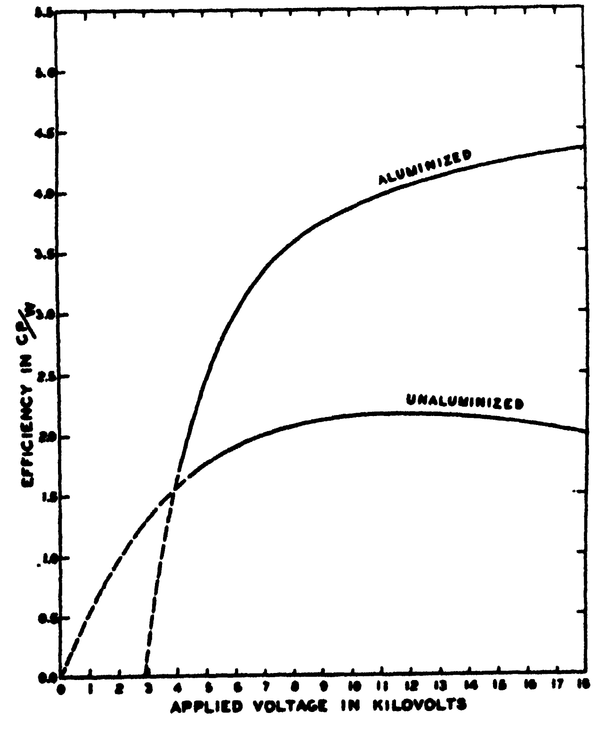
Fig. 7-55.—Fraction of electron beam power passed by aluminum film as a function of beam voltage and film thickness.

that a tube operating at 10,000 volts and with a film about 3,000 Å thick will show no difference from an unaluminized standard tube at the same voltage. It should be noticed, however, that a moderate increase in voltage causes a rapid decrease in percentage loss of beam energy in the film. Experience has indicated that the most useful range of film thickness is between 500 Å and 5,000 Å.

"In order for the film to be relatively smooth and mirror like (condition 2), it has been found possible to cover the fluorescent screen with a thin film of organic material stretched over the crystals like a blanket. This provides a smooth surface upon which the aluminum can be evaporated. If such an intermediate film is not present, the aluminum will be broken up on evaporation so that it will not have its reflecting properties nor will it be continuous and conducting in the thicknesses necessary for low voltage operation. In order to obtain conductivity without the organic film it would be necessary to evaporate five to ten times as much aluminum as is now necessary. This is why previous metallized screen tubes were restricted to high voltage operation.

"These earlier tubes were aluminized in order to avoid undesirable effects due to poor secondary emission from the screen. It can readily be shown that if one tries to operate tubes at a voltage such that the secondary emission ratio from the screen (ratio of secondary electrons to primary electrons) is less than one to one, the screen will accumulate sufficient charge to slow up approaching electrons to a velocity at which the secondary emission is unity. This means that the screen is effectively operating at a voltage that may be considerably less than that applied to the tube. This is known as the "sticking" effect and is almost entirely corrected by providing a conducting layer over the screen. The new method of providing an aluminum film makes possible the correction of the effects due to secondary emission difficulties in tubes operating in the voltage range in which kinescopes are now operated. Thus the choice of luminescent materials for the screen is enlarged and improved techniques for applying these screens to the tube face are made available.

"This aluminum film also provides a new line of attack on the old television tube problem of ion spot. Ions can be completely stopped by a film of aluminum that will readily pass electrons. Experience with tubes in the laboratory has shown that, with the right set of conditions such as proper aluminum thickness and reasonably low gas pressure, tubes can be made which will show no ion spot at normal operating voltages.



Courtesy RCA

Fig. 7-56.—Variation of efficiency with applied voltage of 12-inch tubes, one aluminized.

"Among other advantages of the aluminum film are the protection of the phosphor during processing and life, and the improvement of the stability of the pattern with regard to displacement due to surface leakages on the face of the tube such as are produced if one touches the face of an operating kinescope.

"Fig. 7-56 gives the efficiency in candle power per watt as a function of applied voltage and at a fixed beam current obtained for two laboratory-made 12-inch tubes, identical except that one was aluminized. These curves are typical of measurements made on a number of tubes. It is to be noted that at the lower voltages the unaluminized tube has the higher efficiency whereas above the cross-over voltage the aluminized tube has the higher efficiency. The cross-over voltage which is controlled by the aluminum thickness is primarily dictated by such considerations as operating voltage and ion spot elimination. As seen from Fig. 7-56, the increase in efficiency above the cross-over voltage is quite considerable; for luminescent screens with poor secondary emission characteristics, the gain may be considerably greater than that shown on the figure."

Dark-Faced Tubes

One of the biggest problems in television picture tubes is the attainment of proper brightness and contrast to satisfy the viewer. With ordinary picture tubes, the viewer sees the screen through the glass face of the tube as a light surface having a whitish appearance. In the reproduction of a picture, the electron beam when it strikes the screen causes radiant energy to be released, and hence light output results. This means that those sections of the screen that are hit by the beam produce the light output, whereas the remaining sections of the screen act as background. For a good picture, that is, for proper contrast, the background should be dark compared to the brightened portions. The background can never be darker than the normal appearance of the screen when it is not luminescent.

Since the dark portions or the background of the picture have a light shade which cannot be made darker, the brightness of the illuminated portions of the screen has to be increased to improve the contrast. The need for excellent contrast is pronounced when small details of the picture are to be observed or when there is a high ambient light level which will cause an effective reduction in light output from the tube. Another factor which causes loss of contrast is halation, that is, light which is reflected back to the screen from the inside face of the tube. Special filters of various types have been used in the past to help increase contrast, primarily by eliminating the glare caused by high ambient-light conditions.

If a filter could be designed whereby the loss in contrast due to high ambient light and halation can be diminished, while at the same time darkening the appearance of the face of the tube, we would have an excellent means of increasing the contrast without having to increase the brightness. Today such systems do exist which, though differing in construction and design, all essentially increase the contrast of the tube. Picture tubes that employ any one of these designs are known by various names, among which the "black" tube and "dark-face" tubes are common. Other names, generally trademarks, are applied to specific designs.

One of the first designs used to accomplish this effect was a special glass that possessed many properties desirable for use as the face of the picture tube. Metal-type cathode-ray-tube construction was used, with the exception of the face of the tube. The glass was light absorbing and appeared much darker than ordinary tubes. Therefore, by virtue of these qualities, the high ambient light will be partially absorbed, and halation is reduced, as well as the useful rays producing the desired light output.

It may be asked at this point how beneficial such a glass can be, since some of the useful rays themselves are lost. From the previous discussion on metal-backed screens, we remember that 15 to 25 per cent of the total light output produced is lost due to internal reflections in the glass face (i.e. halation), and that only about 25 to 35 per cent of the total light is useful. Comparing these two figures, we see that the useful light output is only about 10 per cent higher than that causing halation. Now, if we reduce halation by twice as much as we reduce the useful light, the ratio of useful light to halation will be considerably improved, resulting in better contrast. It is easily seen that the light causing halation has to travel through the light-absorbing glass twice, once in the forward direction and then back to the screen (after reflection), while the useful light passes through the glass only once. Hence, the light causing halation is reduced twice as much as the useful light. The result is an over-all effective gain in useful light output without recourse to increased brightness. The dark appearance of the tube face further improves the contrast, because the light portions appear brighter against a darker background.

Another method of improving the contrast employs a special crystalline material, which is very finely coated over the outside face of the tube. This crystalline material is polarized and later covered with a protective coating. As far as is known, this type of tube primarily reduces glare and hence loss of contrast due to ambient light.

The Skiatron

The Scophony Corporation of America has produced a special projection tube called the *Skiatron,* employing a new type of crystalline screen material which produces a *dark trace* when bombarded by electrons. Our interest in this chapter is in the screen material; further information about the tube can be found in Chapter 21.

Some translucent substances, primarily alkali halide crystals, have the property of becoming dark or opaque when bombarded by electrons. When such a crystalline material is bombarded, some of the electrons are displaced from the halides and absorb light at their characteristic frequencies, producing a dark pattern on the originally translucent crystal layer. When a picture is reproduced, it will appear upon the screen the same way as an ordinary film negative. In other words, the final picture will have dark traces against a light background. The greater the number of electrons falling in any one given area, the darker will be that portion of the screen; those portions of the screen not hit by electrons will allow light to pass and hence appear as bright background. If a few electrons fall upon a particular point or section of the screen, this section will take on some shade of gray.

These screens have been primarily designed for projection television systems. An interesting feature of the screens is their memory characteristic, that is, they have a relatively long persistence of the dark trace, which limits contrast and attainable "brightness;" this characteristic makes Skiatrons very useful as PPI radar indicators.

CHAPTER 8

SPOT DISPLACEMENT

Now that we know the function of the different parts of the cathode-ray tube, we are ready to examine the motion of the electron beam in terms of the movement of the luminous spot which it develops on the screen. Up to this point, our discussion of the beam motion was general, being nothing more than an analysis of the deflection forces and the significance of vertical and horizontal motion of the beam.

Inasmuch as the operation of the cathode-ray tube in an oscilloscope or in other types of equipment may require the movement of the spot across all parts of the screen, it is interesting to analyze the relationship between the deflection forces, the spot movement on the screen, and the screen surface. At this time, we are not concerned with the development of any specific type of trace on the screen. That will come later, in subsequent chapters; right now we want to look at the details which describe where the spot can be placed on the screen.

You understand by this time that whatever the nature of the trace on the tube screen, it is the result of the motion of a small luminous spot. You also know that the time allowed for the electron beam to strike a single point on the screen surface determines the luminosity of the spot developed at that point, assuming constant values of operating potential. So, in this chapter, we shall disregard these details and assume a spot of constant density.

The Undeflected Spot

By all normal reasoning, one would expect that when deflection forces are absent in a cathode-ray tube, although the electron beam is present, the normal position of the spot would be the midpoint of the screen, as in Fig. 8-1. If this were a magnetic-type cathode-ray tube, the same normally would be true when current was absent from the deflection coils.

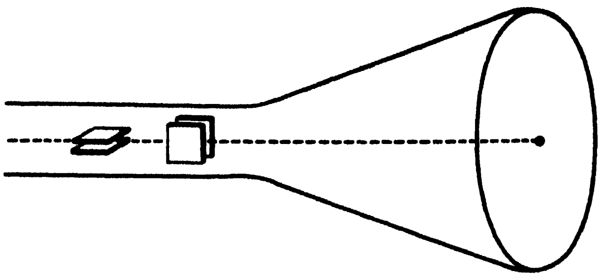


Fig. 8-1.—Ideal position of the spot at the midpoint of the screen; to achieve this usually requires special positioning circuits.

Although this is the desired condition, it can also be viewed as being the ideal condition. By this we mean that, by and large, the undeflected position of the spot *seldom* is at the geometric center of the screen. It must be ***deliberately positioned*** there by some means. The tolerances used in cathode-ray-tube manufacture are of necessity such that satisfactory performance of the gun system is accepted as being *within* a prescribed distance from the exact center of the tube. This was referred to earlier in this text and is repeated here because of other related matters which demand association. So, when we see the undeflected spot at the exact center of the screen, it has been placed there. A centered spot is a rarity without such special positioning circuits; the usual position would be somewhere within a one-inch radius around the screen center.

The fact that the spot must be *positioned* at the center of the screen does not in any way reflect on the quality of the tube. Having once been placed there, it will return to that position when the deflection traces are removed. Of course, minor changes in the operating voltages employed with the tube may slightly affect its exact centering but this is of no consequence, since it can be recentered whenever so desired. As a matter of fact, positioning circuits are a part of every cathode-ray-tube system, so that the spot, or for that matter the trace as a whole, may be located at any point on the screen at the will of the user.

Off-Center Location of the Undeflected Spot

Our comments about the location of the undeflected spot on the screen may imply that such location is imperative to the correct functioning of the tube, or to the use of the tube. Such is not the case, although there is a reason for considering the ideal location as being the geometric center of the tube. It is nothing more than that such positioning affords the maximum useful screen surface area in *all* directions.

It is possible, and not unusual in certain kinds of applications of the cathode-ray tube, that while maxi-

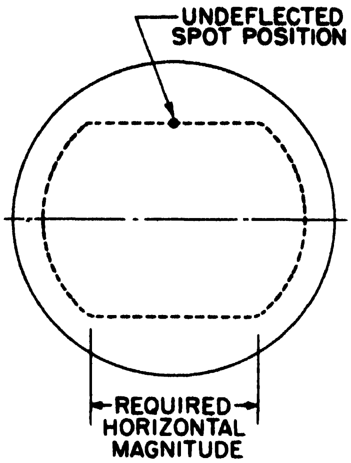


Fig. 8-2.—The dotted lines indicate the useful screen surface area with off-centered undeflected spot position.

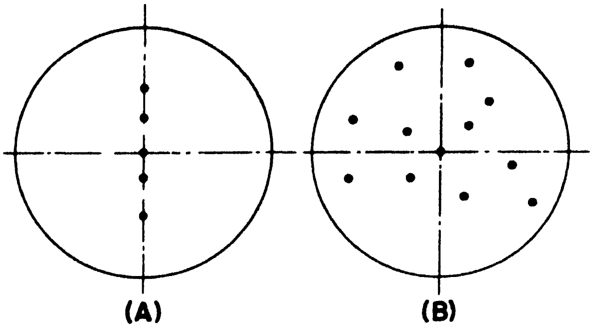


Fig. 8-3.—(A) Straight up-and-down movement of the spot along the vertical axis; (B) random spot positions illustrating vertical deflection while horizontal deflection also exists.

mum screen surface area is desired, it is to be utilized in one direction only, that is, either upward or downward from the undeflected spot position. Such might be quantities which do not appear on both sides of a zero axis, as for example pulses, rectified waves, and the like. In such cases, the best location of the undeflected spot might be toward either the upper limit of the screen as shown in Fig. 8-2, or toward the lower limit.

Of course, any location of the undeflected spot above or below the center of the screen tends to limit the permissible movement of the spot in the horizontal direction because the screen is circular in shape and the longest chord of a circle is its diameter. So it becomes a matter of what horizontal magnitude is required in the display. The dotted lines shown in Fig. 8-2 indicate the useful expanse of the display with such off-centered undeflected spot position. Because of the curvature of the screen near the edges, the useful portion of the screen surface is substantially less than the complete screen area.

As will become evident later, the operation of a cathode-ray tube in such manner that the display is not symmetrical relative to the center of the screen is not predicated on the initial location of the undeflected spot. We have set this primary condition as a basis simply as a matter of convenience. Positioning of the trace as a whole, so that it is most suitable for study or observation, can take place after it has appeared on the screen. This is the equivalent of positioning the undeflected spot, except that it is done after the beam has been deflected.

Spot Displacement in Electrostatic Tubes

Let us now examine the movement of the spot during various conditions of deflection. Having already described the principles of the action, we shall not repeat them; instead we will devote our attention to the excursions of the spot under different conditions. The simplest of these is the use of steady values of d-c voltage as the deflection voltage.

Vertical Deflection

When we say that the luminous spot on the screen moves in a vertical direction from its undeflected position, it can have two meanings. It can mean a straight up-and-down movement along the vertical axis of the screen as shown in Fig. 8-3A, wherein several spot positions other than the undeflected location are indicated. The second meaning, which is related to the first, is *vertical displacement from the horizontal axis of the screen* parallel to the vertical axis. Fig. 8-3A shows this, except that the movement of the spot is *along* the vertical axis. This condition can prevail only if a horizontal-deflection force is absent. If it is present, vertical displacement means the movement of the spot to various points on the screen surface, all of which are either above or below the horizontal axis of the screen, as indicated in Fig. 8-3B. The spot positions shown in this figure are random and purely illustrative of what is meant by vertical deflection while some sort of horizontal deflection also exists.

So far we have spoken in generalities. Let us now become specific and set up an imaginary tube system using an electrostatic type of cathode-ray tube and a variable source of d-c voltage. The voltage will be applied first to the vertical-deflection plates, as shown in Fig. 8-4. Unit *B* is a battery or other d-c voltage source having the polarity shown. Potentiometer *P* is connected across the voltage source, and is the means whereby we can obtain any desired voltage between zero and the maximum rating of the battery. The voltmeter *V* indicates the voltage applied across the plates.

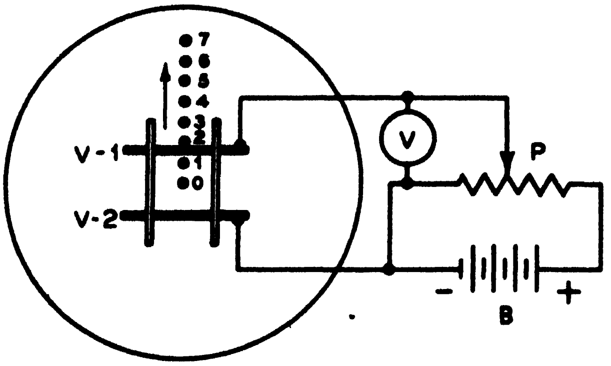


Fig. 8-4.—Adjustment of potentiometer arm *P* for different values of applied voltages displaces the spot from *0* to *7* on the screen.

In this type of circuit, a *fixed* polarity is maintained between the deflection plates, since *V-2* remains connected to the negative terminal of the voltage source. The voltage between these two plates will vary with the position of the slider on *P*, but *V-1* always will be positive relative to *V-2*, except when the applied voltage is zero.

Since the change in voltage applied across the deflection system is ***not*** accompanied by a change in polarity, the ***direction*** of movement of the spot will remain constant, being, in accordance with basic principles, toward the positive plate. The movement of the spot from the undeflected location *0* (midway between the two vertical plates, *V-1* and *V-2*) can be correlated with the exact voltage shown on the voltmeter *V*. This is not essential to the understanding of the action because the specific movement of the spot (the electron beam) and the voltage applied across the deflection system is a function of the tube being used, but it is indicative of one interpretation of spot movement, that of voltage measurement. So that we shall have some correlation with practical operating conditions, let it be assumed that the full voltage of battery *B* is 100 volts. This is a fair approximation of the potential difference required between the vertical-deflection plates in order that a deflection of reasonable proportion be obtained. This should not be confused with the voltage ***which might be required*** at the input of an amplifier, if one were used; it is the voltage at the deflection plates and would conform to the meaning of the tube deflection factor.

As seen in Fig. 8-4, the voltage between the plates is zero when the potentiometer arm is at the extreme left and maximum when it is at the extreme right. The dots bearing numbers from *0* to *7* indicate the relative positions of the spot for different values of applied voltage in arbitrary 10-volt steps. Since we are working with a P1 screen, the persistence of the image is short and only one spot will be visible at one time. If the potentiometer arm is moved very rapidly along the resistor element, even a P1 screen will allow a certain persistence of image, so that what appears like seven individual spots a certain distance apart, might appear as a line. For the sake of clarity, we shall assume a slow movement of the potentiometer arm; moreover, a sufficient movement of this arm is assumed so that the spots appear individually with space between, rather than immediately adjacent to each other. They can be placed adjacent by simply moving the potentiometer arm that small amount which would change the deflection voltage just sufficiently to displace the spot on the screen by an amount equal to its diameter.

Let us start our analysis with the potentiometer arm at its extreme left-hand position. This corresponds to zero voltage, placing the undeflected spot at point *0*. Because *V-1* is positive relative to *V-2*, any change in voltage will cause the spot to move upward, since *V-1* is attracting the electron beam and *V-2* is repelling it. Seven uniform increases of the potentiometer-arm adjustment will shift the spot from position *0* to position *7*. If the potentiometer arm is moved rapidly, the difference of potential between the plates will be changed rapidly and the spot will move in synchronism with the change in voltage. If the change in voltage is sufficiently rapid, the eye will not see the spot in motion, but will see the spot trace a luminous line. If, during this operation, the potentiometer arm is stopped anywhere between its limits, the developing line will suddenly become a single spot of small dimensions, corresponding to the static value of voltage established by the setting of the potentiometer arm.

With the electrical system as shown, one limit of the spot excursion is the undeflected position *0*, conforming to zero voltage from the voltage source; the other is position *7*, which conforms to the maximum voltage available from the voltage source. At no time can the spot move any closer to *V-2* than position *0*, because *V-2* cannot be made positive relative to *V-1*. Thus it is evident that a d-c voltage will move the spot in one direction only, unless the polarity of the voltage is reversed. Another highly significant point is that the magnitude of the spot displacement is proportional to the value of deflection voltage. For example, if the vertical excursion of the spot between *0* and *7* in Fig. 8-4 is 1.5 inches, and the range of voltage shown on the voltmeter is from 0 to 100 volts, then a displacement of spot from the undeflected point *0* to point *7* corresponds to a deflection voltage of 100 volts. Then each indicated point between *0* and *7* corresponds to a voltage increment of 100/7, or 14.28 volts.

A displacement of the spot from *0* to *1* means a deflection voltage of 14.28 volts is applied; from *0* to *2* it means a deflection voltage of 28.56 volts, etc. These voltage references are not intended to show the actual application of the cathode-ray tube as a voltmeter; it is entirely too soon in this text to introduce this subject. All we have in mind at this time is to direct your thinking toward certain applications to be mentioned later. One of these is voltage measurement; another is the fact that if a change in the deflection voltage is linear, the motion of the spot on the screen will be linear, assuming, of course, that there are not other influences which might tend to alter the normal behavior of the tube system.

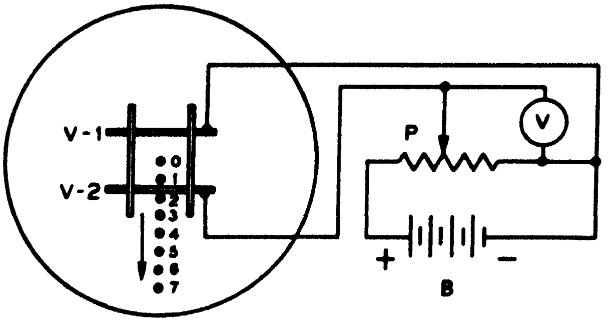


Fig. 8-5.—Vertical spot displacement with reversed polarity of the deflection plates; *V-2* positive, *V-1* negative.

Suppose that we reverse the deflection voltage by reversing the polarity of the voltage source, as in Fig. 8-5. The system is the same as before, except that the polarity of the deflection plates *V-1* and *V-2* now is the reverse of that shown in Fig. 8-4. Now we see that the spot moves toward *V-2*, which is natural, since it is positive relative to *V-1*. Since the maximum range of potential is the same as before, the spot traverses the same distance across the screen for the same swing of the potentiometer arm. This indicates that as far as the tube is concerned, it does not distinguish between polarity of deflection; it is strictly the voltage across the deflection system which controls the amount of electron-beam deflection. We might restate this by saying that the movement of the spot is linear for linear changes in voltage, regardless of the instantaneous polarity. This is indicated by the separation between points *0* and *7* in Fig. 8-5 being the same as that between the two corresponding points in Fig. 8-4. So, as in the upward deflection, each point between *0* and *7* equals a deflection voltage of 14.28 volts, for, after all, it is still the same tube and the elements were not changed; all that was changed was the polarity.

Still another very important point which might be mentioned is an interpretation of applied polarity by observation of the direction of spot movement. Suppose, for example, it were known that the orientation, relative to the horizontal axis, of the vertical-deflection plates of the tube was such that a positive voltage applied to the upper plate would move the beam upward. The movement of the spot on the screen would then instantly disclose the polarity of the voltage source connected to the deflection plates.

While only seven spot positions are shown here, the possible number of such positions is determined solely by the increment of change of the potentiometer arm. Given sufficient voltage at *B*, the points of maximum displacement of the spot in either direction may be at the extreme limits of the screen — or even completely off the screen. This is possible without having the electron beam actually strike the deflection-plate surface and thus be impeded in its progress toward the screen. With sufficient deflection-voltage amplitude, the angle at which the electron beam is tilted before striking the deflection plate may be sufficient to cause it to strike the wall of the envelope, well off the screen.

Although previously implied, it must be understood that the vacant positions between the spots in Figs. 8-4 and 8-5 are not normal; they are strictly arbitrary and brought about by the fact that we temporarily ignore the motion of the beam between the indicated positions of rest. Actually, the beam does not *jump* from one position to the other unless the source of voltage responsible for the deflection force is of that character which changes voltage in *abrupt* steps — and shuts off the beam during the moment of change. Even with abrupt changes of deflection voltage, as in the case of square waves, the movement of the spot will be discernible unless the change is extremely rapid.

Since the entire purpose of the equipment is to display information on the screen, it stands to reason that a voltage which will cause the spot to move off the screen is representative of excessive deflection voltage. As a matter of fact, it is appropriate at this time to recall that the useful limit of the screen surface is not the periphery. It is somewhat less than this limit all around the screen surface. For proper presentation, that is, minimum distortion of the spot, approximately one-quarter to one-half inch from the extreme edges of the screen should be the maximum limits of deflection in all directions.

Horizontal Deflection

What is true about the vertical-deflection plates applies to the horizontal-deflection plates. As shown in Fig. 8-6, the voltage source is the same as before and everything else is the same except that the horizontal plates are identified as *H-1* and *H-2*. With *H-1* positive with respect to *H-2*, the beam is attracted by the former and the spot moves in the direction of the arrow. The exact position of the spot is determined by the potential difference, as previously described. If we reverse the polarity of the voltages applied to the two horizontal-deflection plates, as in Fig. 8-7, wherein *H-2* is positive and *H-1* is negative, the beam will be attracted toward *H-2*, and the spot will move toward the left.

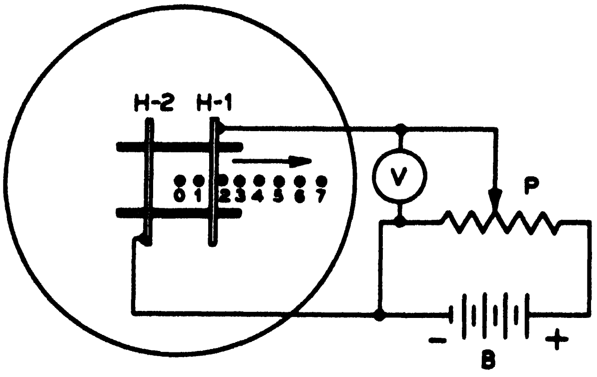


Fig. 8-6.—Horizontal spot displacement with *H-1* positive with respect to *H-2*.

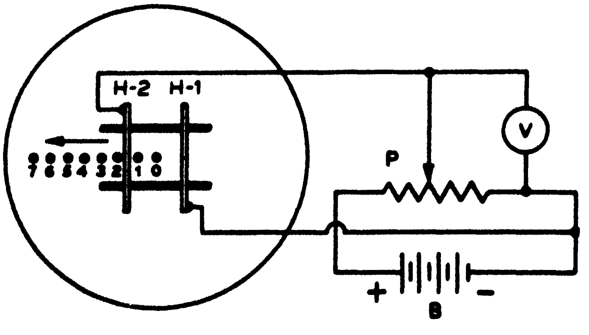


Fig. 8-7.—Horizontal spot displacement with reversed polarity; *H-2* positive, *H-1* negative.

Bilateral Deflection

Let us now rearrange the voltage source so that we can conveniently change the polarity of the voltage applied to the deflection systems; the circuit is shown in Fig. 8-8. The battery and potentiometer system is modified so that by moving the arm of potentiometer *P*, the polarity of the voltage applied to the two deflection plates can be reversed. With the potentiometer arm set to the midpoint and the battery tapped at its midpoint, the potential difference between the two plates is zero; hence the spot occupies the position designated as *0*. If the arm is moved to the extreme left position, plate *V-1* is made maximum negative with respect to *V-2*, and the spot moves toward *V-2*, eventually reaching position *7'*. If the potentiometer arm is moved to the extreme right, *V-1* is made positive with respect to *V-2* and the spot moves toward *V-1*, eventually reaching position *7*. Therefore, by moving the potentiometer arm between its limits, we can cause the spot to move vertically across the viewing screen to whatever limit is set by the voltage difference existing between the two plates, this being determined by the stationary position of the potentiometer arm.

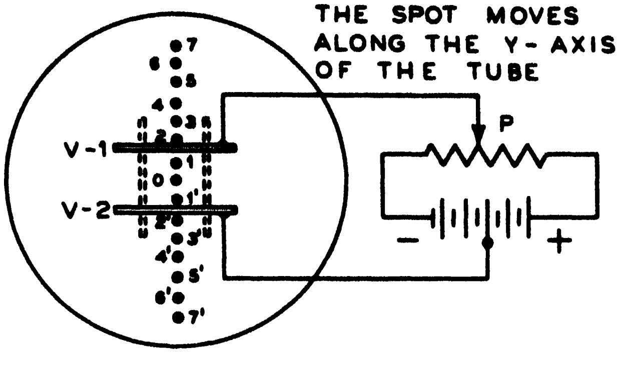


Fig. 8-8.—Vertical spot movement up and down from center position; if potentiometer arm is moved left from center, *V-1* becomes negative with respect to *V-2* and the spot moves downward; if the arm is moved toward right, *V-1* becomes positive with respect to *V-2*, and the spot moves upward.

If the potentiometer arm were moved from its midpoint toward the right, then reversed to the left, and finally brought back to the midpoint, the spot would move upward from *0* to *7*, back down to *0*, continue downward to *7'* and then return to *0*, where the arm is again brought to rest at the midpoint on the potentiometer resistance.

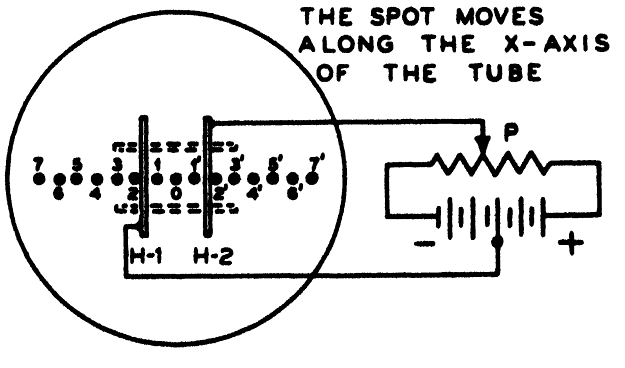


Fig. 8-9.—Horizontal spot movement to the left and right from center position; if potentiometer arm is moved left from its center position, the spot moves to the left; if the arm is moved right from center, the spot moves toward the right.

That which is true about the action of the vertical-deflection plates also applies to the horizontal plates, as shown by the positions of the dots in Fig. 8-9. Once more we repeat that the seven dots, representing spot locations each side of the undeflected beam position, do not establish that the total number of spot positions is seven each side of *0*. These are arbitrary selections of spot movement along the horizontal axis.

Resultant Spot Deflection

Let us now consider the possible range of spot positioning, with respect to the complete viewing screen. Up to the present time, we have considered but one deflection force applied to the beam which passes between the set of active deflection plates. From now on we shall work with two deflection voltages, one voltage applied to one set of plates and another voltage applied to the other set of plates. As before, we are employing d-c voltages secured from batteries.

Suppose that we start with Fig. 8-10A. Here we see two voltage sources, B_V and B_H, controlled by two potentiometers, P_V and P_H. Once more the battery circuits are such as to cause unilateral deflection in each deflection system. If we look at each of these separately (neglecting the presence of the other), the influence of the voltage applied to the vertical-deflection plates would cause the spot to move toward *V-1* and occupy positions as indicated by the circles marked *1V, 2V, 3V, 4V,* etc. If we consider the horizontal deflection separately, the positions of the spot would be in the direction of the arrow *H*, at the circles *1H* to *7H*.

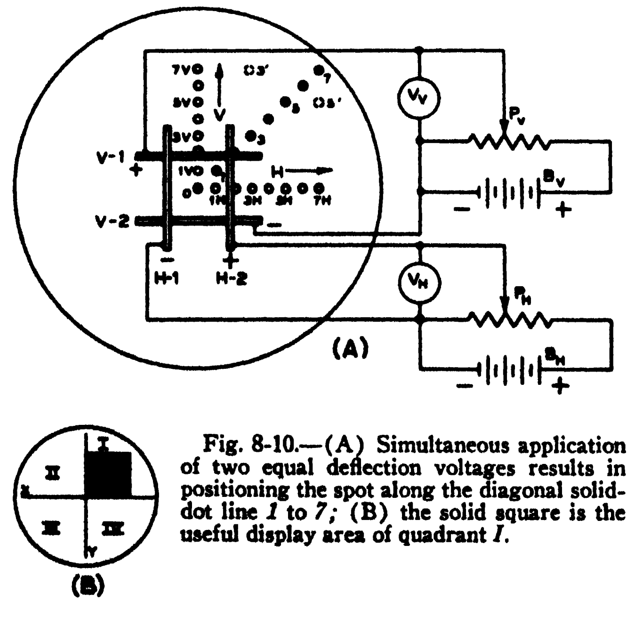


Fig. 8-10.—(A) Simultaneous application of two equal deflection voltages results in positioning the spot along the diagonal solid-dot line *1* to *7*; (B) the solid square is the useful display area of quadrant *I*.

However, both batteries are applying deflection voltages simultaneously, which act upon the beam at right angles to each other; consequently, the final position of the spot is determined by *two* deflection forces acting at right angles to each other. If the voltage applied to the vertical plates is such as to move the beam to the position *1V*, (with B_H inactive), and the voltage applied to the horizontal-deflection system is such that it would move the electron beam to create a spot at *1H* (with B_V inactive), the combined force produces a *resultant* position of the spot indicated by the solid dot *1*. If you examine the location of this dot, you will find that displacement along the *Y*-axis, that is, the vertical axis, is the same as for the vertical deflection, without any horizontal deflection, and that the displacement from its basic position along the *X*-axis, that is, the horizontal axis, is the same as for the horizontal deflection, without any vertical deflection.

If the individual voltages, with the polarities shown, are such as to cause the individual deflections corresponding to *2V* and *2H*, the resultant deflection sets the spot at the position shown by the solid dot *2*. In like manner, the other final locations of the spot (solid dots *3* to *7*) are developed by the simultaneous application of two deflection voltages, each of which individually would cause the spot to occupy the points shown by the vertical and horizontal rows of circles.

In the example considered, the magnitudes of the individual vertical- and horizontal-deflection voltages are deliberately made equal. In practice this may or may not be so, but it lends itself best to the basic presentation. A typical practical case might be simultaneous vertical- and horizontal-deflection forces corresponding to *7V* and *3H*, which would then move the solid dot *3* from its position for equal voltages to the point indicated by the dotted circle *3'*.

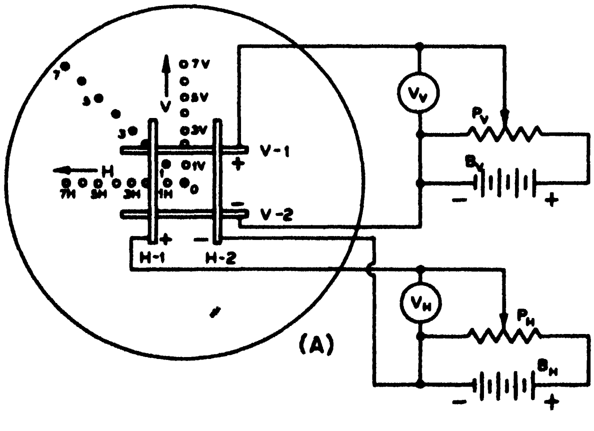
Another example of this might be a vertical-deflection voltage corresponding to circle *5V* and a horizontal-deflection voltage corresponding to circle *7H*, in which case the spot *5* would be at point *5'*, indicated by the dotted circle.

Various values of vertical- and horizontal-deflection forces due to different values of unidirectional voltages can cause the ultimate positioning of the luminous spot at any point in the upper right-hand quarter of the screen. Technically such a part of a circle is known as a *quadrant*, meaning one-fourth of the area of a circle. The area indicated by the solid square in Fig. 8-10B is the useful part of quadrant *I*. Actually the entire quarter section of the screen is the quadrant, but since only a part of the screen can serve as a useful display area, we feel that it is best to most closely associate utility with the division of the screen area. As you can see, the entire screen surface is divided into four quadrants — *I, II, III,* and *IV.*

Relative to the representation of the possible positions of the spot in the useful area of quadrant *I,* illustrated in Fig. 8-10B, it is interesting to examine why the boundaries of this area are as they are shown. In the fewest words it is due to: (a) the use of unidirectional deflection voltages for both the vertical and horizontal deflections, and (b) the fact that zero deflection voltages on one set of plates locate the spot along the axis related to the other set of plates. With the deflection voltages applied as shown, the spot cannot move to the left of the vertical axis or below the horizontal axis.

Since maximum vertical-deflection voltage and zero horizontal-deflection voltage position the spot along the vertical axis, that axis becomes one boundary of the useful spot area in quadrant *I.* Another boundary of the solid square lies along the horizontal axis, because with different horizontal-deflection voltages and zero vertical-deflection voltage, the spot cannot move below the horizontal axis. The other two boundaries of the solid square in Fig. 8-10B are set by the maximum vertical- and horizontal-deflection voltages. The solid dots form one diagonal of the square. Dotted circles *3'* and *5'* indicate the development of the upper and right-hand boundaries of the solid square of dot positions by the application of unequal values of vertical- and horizontal-deflection voltages.

If you now refer to Fig. 8-11A and examine the polarity of the voltages applied to the plates, you will find that the upper vertical-deflection plate *V-1* is still positive. However, the polarity of the horizontal-deflection plate system has been revised. Now *H-1* is positive and *H-2* is negative. Displacement of the beam spot takes place exactly as before, but in this instance, the horizontal motion is toward the left horizontal-deflection plate; the resultant of the vertical- and horizontal-deflection forces places the active area of the spot position in quadrant *II,* as shown in Fig. 8-11B. The forces acting upon the beam are the same in Figs. 8-10A and 8-11A, this being illustrated by the use of identical arbitrary units upon the drawing. It is the change in the direction of the horizontal deflection which is responsible for locating the spot position to the left of the vertical axis.



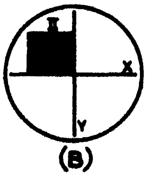
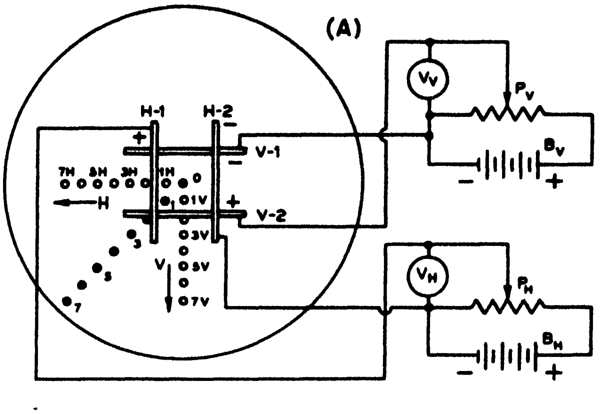


Fig. 8-11.—(A) With reversed polarity, *H-1* positive and *H-2* negative, the spot position area is in quadrant *II,* (B) the active area of quadrant *II* is indicated by the solid square.



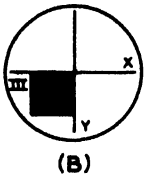
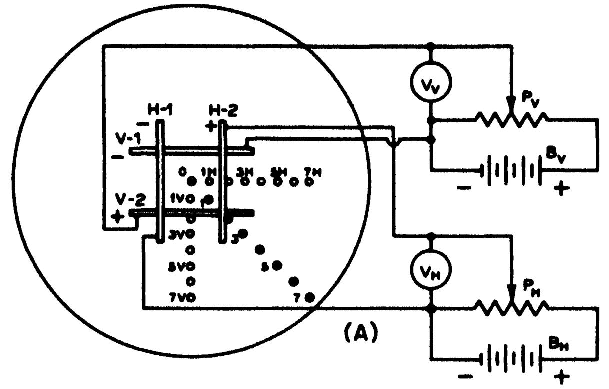


Fig. 8-12.—(A) Changing the polarity of the vertical-deflection voltage and retaining the polarity of the horizontal-deflection voltage locates the spot in quadrant *III,* (B) the active area is indicated by the solid pattern.

By changing the polarity of the vertical-deflection voltage and retaining the polarity of the horizontal-deflection voltage as shown in Fig. 8-11A, we can position the beam spot *below* the horizontal or *X*-axis. This is shown in Fig. 8-12, where by making *H-1* and *V-1* positive by varying degrees relative to *H-2* and *V-2,* the spot is positioned to the left of the *Y*-axis and below the *X*-axis. This locates the spot in quadrant *III,* and the active area is indicated by the solid pattern in Fig. 8-12B. The action of the deflection forces is identical to that already explained, except for the change in the polarities of the deflection systems.

Placement of the spot in quadrant *IV* is shown in Figs. 8-13A and B. As far as the action involved is concerned, it already has been described and requires no repetition. That which was described in connection with the basic illustration, Fig. 8-10, is applicable in



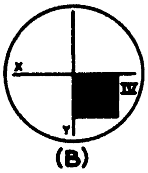
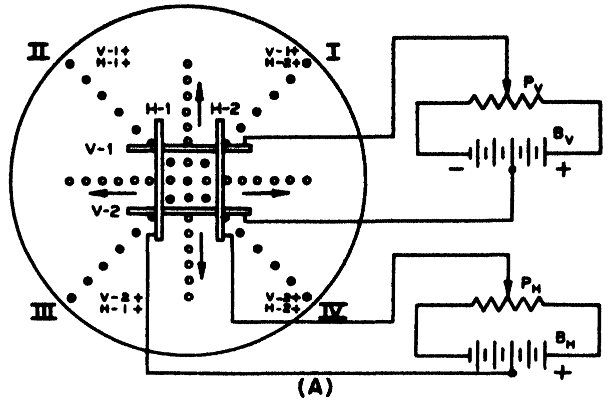


Fig. 8-13.—Placement of the spot in quadrant *IV*.



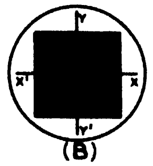


Fig. 8-14.—(A) The basic vertical- and horizontal-deflection spot positions are indicated by the circles; the solid dots indicate the spot positions when both the horizontal- and vertical-deflection voltages are varied at the same time and have the same magnitude; (B) for all possible values of voltage applied to the two sets of plates, the spot is within the solid square.

its entirety. However, in the present instance, in any one of the other cases, positioning of the spot in any given quadrant is a matter of applying to the deflection system voltages of the correct polarity.

Summarizing all that has been illustrated relative to the four quadrants of the cathode-ray-tube screen:

1. Changes in *vertical* deflection-voltage polarity place the active area *above* or *below* the *horizontal axis.*
2. Changes in *horizontal* deflection-voltage polarity place the active area to the *right* or to the *left* of the *vertical axis.*

In this connection, we must emphasize one very significant point. The association between the specific identifications of the plates in the deflection systems and the quadrant where the spot will appear is an arbitrary one. Someone else may assign different numerical designations to each of the plates in the two deflection systems. This may change the references to the direction of spot movement or quadrants and the *deflection-plate identification;* but it will not alter the relationship between the *direction of deflection* and the *polarity* of the deflection plates. The specific numerical identification of a deflection plate is of secondary importance. The important point is the relative polarity of the plates in each of the deflection systems.

Consolidating Figs. 8-10 through 8-14 results in Figs. 8-14A and B. Once more we have arranged voltage sources which afford a complete change in polarity for both sets of deflection plates as the two potentiometer arms are varied along their respective resistance elements. The positions of each, as they are illustrated, produce zero potential difference between plates of each pair. The spot, therefore, is at the exact center of the square formed by the two sets of intersecting plates, and is shown by a solid dot.

Radiating outward from this center position are the basic vertical- and horizontal-deflection spot positions, these being indicated by the circles. These show the movement of the beam in each of the four directions from the center, when the two voltage sources are individually varied over their full range of voltage and polarity. When both voltage sources feed deflection voltages to the two systems simultaneously, and these are of like magnitude, the resultant spot positions are as shown by the solid dots.

The relative polarities of the plates for deflection in each quadrant are indicated in terms of the vertical plate and the horizontal plate which must be positive. Thus, in the area which is the equivalent of quadrant *I,* vertical-deflection plate *V-1* is positive, and horizontal-deflection plate *H-2* is positive. In quadrant *II,* the positively charged plates are *V-1* and *H-1.* In quadrant *III,* they are *V-2* and *H-1,* and in quadrant *IV,* they are *V-2* and *H-2.*

For all values of voltage applied to the two sets of plates, the spot would fall somewhere within the solid square of Fig. 8-14B. This assumes that the magnitude of the deflection voltages is kept within certain bounds as indicated by the limits of deflection set by the extreme positions of the solid dots in Fig. 8-14A. These correspond to the four corners of the solid square shown in Fig. 8-14B. The remainder of the square is made up of dot positions dependent upon different unequal values of horizontal- and vertical-deflection voltage. This is indicated more clearly in Fig. 8-15.

In this illustration, the useful area of the screen is occupied in part by a number of dot positions. Here are shown a number of combinations of vertical and horizontal voltages which will produce spot positions, and which, when illustrated in composite form, will comprise a square. Whereas before, we illustrated equal deflection-voltage magnitudes in both directions, here are shown unequal values of voltage active in the horizontal and vertical directions.

For example, the solid dot in quadrant *I,* identified as *7V, 1H,* is the spot position for 7 units of vertical-deflection voltage and 1 unit of horizontal-deflection voltage. The spot position in quadrant *II* identified as *4V, 7H* shows the spot position when 4 units of vertical-deflection voltage and 7 units of horizontal-deflection voltage are applied. The space between the dots or spot positions is due to the use of 1-volt *changes* in deflection-voltage values. The smaller the increments of these voltage changes, the less will be the separation between the spot positions. This is illustrated by the dot corresponding to a vertical-deflection voltage of 3½ units and a horizontal-deflection voltage of 5½ units. Note its location relative to the neighboring unit-voltage spot positions. If the voltage increments were ¼ units, the spots would be even closer together.

If the changes in deflection voltage, as well as the changes in polarity, were made continuous, the entire active area of the screen would be explored by a moving spot. This gives rise to a subject already mentioned in connection with screens, but repeated here because of its close association with the subject at hand. We are referring to the optical illusion created by a spot of light which is in motion along the same path.

Persistence of Vision

Have you ever seen a flaming torch which was moved in a circular path? To the observer, such a moving point of fire traces out a circle, although actually at any one instant, it is located at a specific point. By retracing its path at a sufficiently high speed, the moving torch deceives the eye; the eye sees a continuous path of light because it cannot separate the individual instantaneous locations of the torch. In like manner, assuming the presence of the proper mechanism for the application of deflection voltages of proper magnitude and polarity, the pattern of Fig. 8-15 can be made to appear like that shown in Fig. 8-16, wherein are illustrated a series of lines which correspond to the horizontal rows of dots shown in Fig. 8-15.

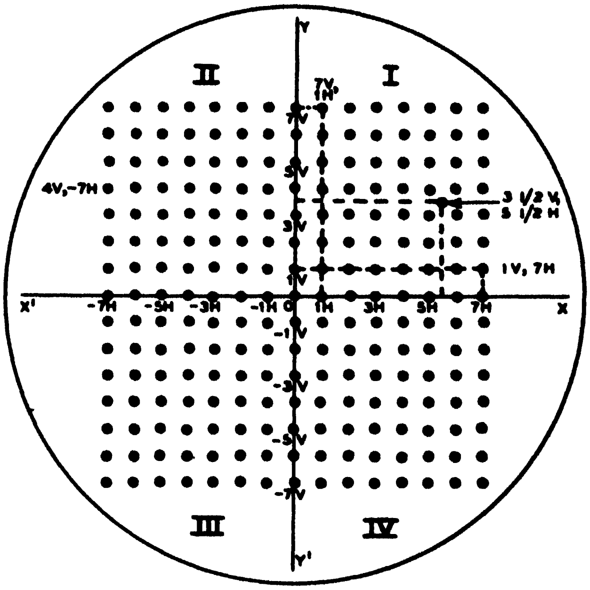


Fig. 8-15.—The dots show spot positions for a variety of combinations of unequal horizontal- and vertical-deflection voltages corresponding to different settings of the two potentiometers, the square indicates useful screen area.

If the increment of change in the vertical-deflection voltage were made sufficiently small, these parallel lines would seem to touch each other; in fact, they could not be distinguished from each other. If the sequence of deflection voltage and polarity changes were sufficiently rapid, and the spot retraced its path time and again, the eye would be fooled completely. Although a single spot is doing the tracing of the pattern, the screen would appear to the eye as an illuminated square. This is why we can think in terms of solid patterns of light formed by different positions of the luminous spot, as was done in Figs. 8-10 through 8-14.

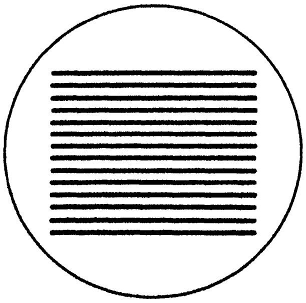


Fig. 8-16.—Series of lines seen due to persistence of vision which correspond to the horizontal rows of dots shown in Fig. 8-15.

The phenomenon is not limited to horizontal lines as shown in Fig. 8-15. It is just as applicable to vertical lines, or to lines in any direction. It is simply a matter of the path being traced out by the moving spot of light. The eye is capable of distinguishing steps in motion, if they occur slower than about 16 times per second. At higher speeds, it *appears* as continuous motion. This is the basis of motion pictures. These are flashed on the screen at the rate of 24 still pictures per second, and the eye sees them in continuous motion.

The picture which appears on a television picture tube is formed by a spot of light of varying intensity which moves horizontally across the screen. Each excursion is separated from the other by a tiny distance, advancing downward from the top. The picture seen on the screen is traced out by about 490 horizontal lines which appear on the screen 30 times per second. The structure of the picture in the form of lines can be discerned on any television screen which has been adjusted for a sharply focused image.

Spot Displacement in Electromagnetic Tubes

That which has been described as the movement of the spot across a cathode-ray-tube screen under the influence of electrostatic deflection fields, is repeated in magnetic-type tubes when the deflection field is electromagnetic in character. If we are to note a difference, it is found in what has already been described in detail, namely that the movement of the electron beam is at right angles to both the lines of flux and the direction of the deflection force.

In all other respects, the behavior is the same. The electrostatic field set up between the deflection plates, when a voltage is applied, is the counterpart of the electromagnetic field set up between the deflection coils when current flows in the windings. The polarity of the electrostatic system has its equivalent in the direction of the current through the coils, which in turn determines the direction of the flux lines through which the electron beam passes.

The similarity in the resultant spot motion on the screen, in both the electromagnetic and the electrostatic systems, is such as to permit the use of the same illustration for spot positions. Instead of voltage as used in the electrostatic system, we think in terms of current in the magnetic system; but the action of each pair of coils with current in one direction is exactly the same as with the individual sets of plates with a fixed polarity relationship. Thus, the composite illustration of Fig. 8-14A is repeated as Fig. 8-17, with the exception of deflection coils which are used in place of deflection plates. The vertical-deflection coils are designated as V_c and the horizontal-deflection coils are marked H_c.

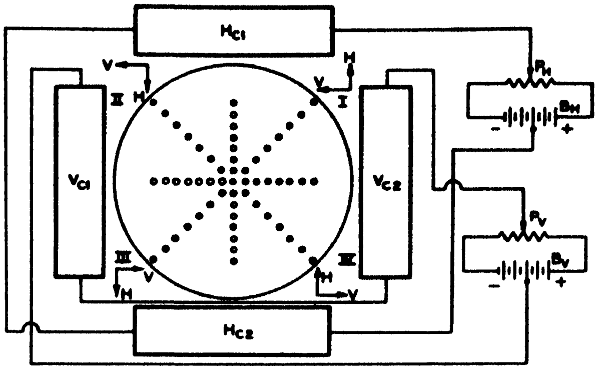


Fig. 8-17.—Deflection coils V_{C1}, V_{C2}, H_{C1}, and H_{C2} are used to develop the spot positions in any one of the four quadrants.

The current sources are shown as batteries with a potentiometer connected across each, so arranged that polarity may be changed, as well as the amplitude of the voltage across the coil, thereby changing the current through the windings. Such sources of current for the coils are not the most practical, but they will serve for this explanatory illustration. The organization of screen quadrants is the same as before, except that in this case, the correlation between the active quadrant and the movement of the electron beam or spot is in terms of the directions of the lines of flux between the vertical-deflection and between the horizontal-deflection coils. The vectors labelled *V* and *H,* near each quadrant designation, show the directions of these flux lines for beam deflection into that quadrant.

It is important that you understand the use of these deflection coils in pairs at all times. Coils V_{c1} and V_{c2} are used to develop the vertical spot positions both above and below the horizontal axis of the screen regardless of which quadrant of the tube is active. The same applies for the horizontal-deflection coils, H_{c1} and H_{c2}, relative to the development of the horizontal spot positions in any one of the four quadrants.

The fact that unilateral deflection can be obtained in any direction is not due to the use of a single coil in either pair, or in both pairs, but rather to the single direction of the magnetic deflection fields. This is brought about by unidirectional current of variable magnitude in the deflection coils.

So, if we correlate the behavior of the system shown in Fig. 8-16 with the electrostatic process, a change in the *direction* of the current through the deflection coils and the change in the *amplitude* of these currents is the equivalent of a change in the voltage applied to the deflection plates and the change in relative polarity of the plates. In this manner, the spot can be moved to any position on the screen surface, and Fig. 8-15 applies equally to electrostatic- or electromagnetic-deflection systems.

Multigun Tubes

It might be well at this time to make a few remarks about the multigun type of tube. The behavior of each gun is the same as if no other gun were in the tube envelope. Nevertheless, certain comments, based on experience, may prove beneficial to the reader.

There is, for example, the matter of undeflected spot position. In the single-gun tube this poses no problem. In the multigun tube, it is not an equally simple matter. In the single-gun tube, it can be centered, if that position is necessary. In the multigun tube, undeflected centering of the electron beam from each gun so that the spots are coincident on the screen is just one of a number of related problems. Normal manufacturing tolerances allow undeflected spot centering to mean any place within a one-inch square around the geometric center of the screen. Adjusting the spots due to two or more electron beams in a tube to be coincident on a single point on the screen is simple; but whether these beams will *maintain* their *parallelism* while being deflected by the same voltage is an entirely different matter.

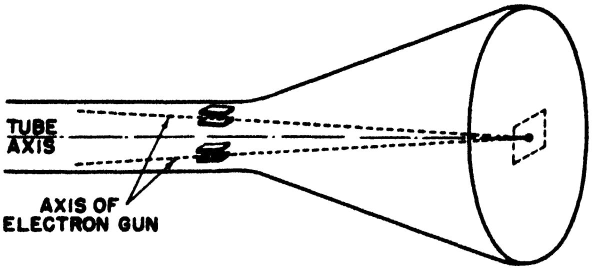


Fig. 8-18.—The multigun tube uses tilted guns and associated deflection systems to position the spot within a one-inch square around the geometric center of the screen.

The usual multigun tube makes use of tilted guns and associated deflection systems. An example of this is shown in Fig. 8-18, which is a rough illustration of the relative tilt of the guns in a dual-gun tube. The deflection systems which are supported by the gun mounting are likewise tilted relative to the tube axis so as to permit the beam to traverse the space between the deflection plates at the center. By tilting the guns in this manner, an attempt is made to attain coincidence of the beams at one point on the screen with minimum correction; at the same time, the greatest screen surface is afforded each system in all directions.

It would be possible to arrange the guns so that their axes were parallel to the tube axis, displacing each by the same amount from the axis of the tube. In this event, the undeflected spots would lie along the vertical axis of the screen, equidistant from the horizontal axis, as shown in Fig. 8-19. At first glance, one might say that this type of mechanical arrangement would tend to limit the magnitude of the display, since the undeflected spots are not in the center. Such is not the case, because the usual design of the deflection systems permits correction of the undeflected spot position in a manner which affords the same space over the screen surface for display as exists with the centrally located spots.

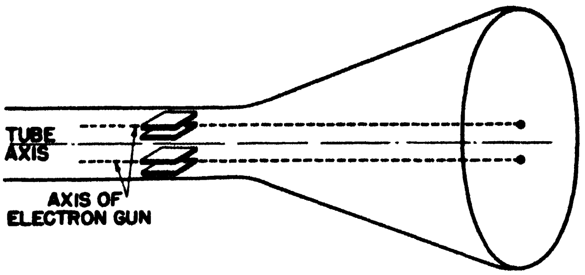


Fig. 8-19.—With the axes of the guns parallel to the tube axis, the undeflected spots lie along the vertical axis of the screen equidistant from the horizontal axis.

Defects in Deflection

The problem associated with the two types of gun arrangements is not really the undeflected spot centering. Of far greater importance is the behavior of the two or more beams — depending on the number of guns — when deflection voltages are simultaneously applied to both deflection systems. There is no difficulty in centering two differently positioned undeflected spots so that they are coincident. The complication is the lack of parallelism when, for example, two electron beams are deflected by identical forces. The greater the number of beams, the greater is this problem.

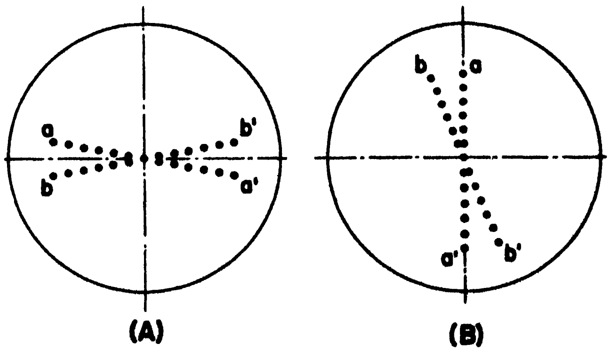


Fig. 8-20.—(A) When two beams are deflected by equal values of voltage, the two traces are tilted with respect to each other; (B) here one of the beams tilted with respect to the vertical axis.

The spot positions of a tube containing two guns can be seen in Fig. 8-20A. Coincidence of the two undeflected beams on the screen is obtained; but when these are deflected by equal values of voltage, the movement of the two beams along the horizontal axis lacks parallelism. Such beam behavior would tend to tilt the two traces with respect to each other. This does not limit ordinary waveform comparison, but it would be a limitation in connection with phase comparisons. The condition we show is somewhat exaggerated; it may be found in actual practice, although perhaps not to this great extent. However, this effect must be reduced quite substantially in order that the phase shift it causes be negligible.

Correction for such a condition is possible, but quite difficult, since it is necessary to compensate for the beam tilt as a whole. Such correction will be discussed later. Sometimes only one of the beams is tilted in this manner, the other beam moving along the horizontal axis. This simplifies the correction and reduces the error. Then lack of parallelism may be found to take place along the vertical axis, as illustrated in Fig. 8-20B. Here the spot positions coincide when the beams are not deflected, but one moves away from the vertical axis when both are deflected.

We are anxious to reiterate that such conditions do not restrict the usefulness of the dual-beam tube. In the case of phase measurements, it simply makes it a little more difficult to interpret that which is seen. As is evident, the minimum effects due to such conditions exist near the center of the tube screen, so that.if the shift due to the lack of parallelism is determined beforehand, the correction factor can be applied. Doubtless, in time, a greater degree of accuracy will be attained in the manufacture of the guns and deflection systems, as well as their orientation relative to the tube axis, and the required parallelism during deflection will be accomplished. Even today, the amount of such error is quite small and our reason for mentioning these details is merely to indicate their existence.

Still another condition sometimes found in multigun tubes is that the deflection sensitivities of the two guns are not the same. The net result is that while parallelism of beam motion and spot movement may be accomplished during deflection, the *amount* of deflection per unit voltage applied across the two like deflection systems will not be the same. What we mean is shown in Fig. 8-21. Here we have deliberately displaced the two spots vertically by a definite amount so as best to show the action.

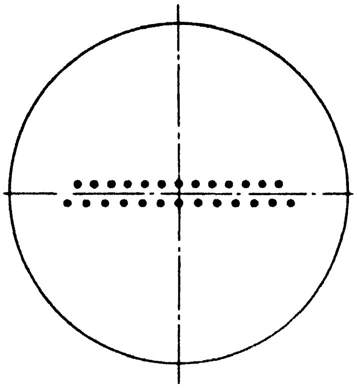


Fig. 8-21.—Amount of resultant deflection per unit voltage applied across the two deflection systems is unequal, because of slight differences in deflection sensitivities of the two guns.

The two spots are lined up along the vertical axis for the undeflected conditions, but, as can be seen, equal values of voltage applied to each pair of horizontal-deflection plates do not move the electron beams and associated spots the same amount. Certain corrections may be applied, but they do not always fully remedy the condition. They may correct it sufficiently so that those tests which might be interfered with most by such nonlinear deflection may be carried out with negligible error. Naturally, the greater the number of guns in the tube, the more difficult is the solution of any one of the nonuniformities described in connection with the dual-gun tube, especially the effort to make all the beams move in perfect synchronism. Nevertheless, these tubes are extremely valuable and are generally more versatile than two single-gun tubes.

Reaction Between Beams

If we continue the assumption that the deflection voltage is applied directly to the plates of the tubes (without any amplifiers in the circuit), we can, for all practical purposes, consider that each electron beam behaves as if no other beam were in the tube. Sufficiently good shielding exists between the deflection plates of each system, so that there is no reaction between the beams. Under such conditions, each beam can be deflected in whatever manner is consistent with proper performance, and the resultant traces on the screen can cross each other as frequently as may be necessary. Or the two beams may be made to produce a coincident trace, without one interfering with the other; each will produce its own trace just as if the other beam were not there.

It is imperative, however, to remark that such freedom from interaction may not be experienced when amplifying systems are used with each of the guns which produce these beams. This does not mean that the addition of the amplifiers feeding the deflection systems causes the beams to interact, but rather that if interaction is noted, it can be attributed to coupling *between the amplifiers or whatever supplementary equipment may be used with the guns and the deflection systems.*

Interpretation of Spot Displacement

Although the displacement of the spot from the undeflected position is the process of tracing a visible pattern of a varying quantity or some other action, we can place another interpretation on the displacement. This is the magnitude of displacement relative to the voltage responsible for it. Restated, the displacement of luminous spot on the beam from any point which is considered the reference point — such as the undeflected location at the center of the tube screen — is a measure of the magnitude of the applied deflection voltage.

This was evident in Figs. 8-4 through 8-15, although it was not discussed to any great length. In each of these illustrations, regardless of the direction of the deflection, the movement of the spot from the center to some point away from there was accomplished by the application of a d-c voltage of some value. Whereas during the discussion of the action we paid very little attention to the relationship between voltage level and distance of spot displacement, we now can bring up that point. Not that we intend to review the past discussions; rather, we desire to emphasize the utility of the cathode-ray tube as a *voltage-measuring device.*

In this connection, it is significant to note that, under normal conditions, the deflection due to a voltage is a linear one; that is, the deflection caused by a given value of deflection voltage will result in equal displacement either side of the undeflected position along any *one* axis, provided that the polarity of the voltage is changed to swing the electron beam both sides of the undeflected spot position. However, it is also significant to note that while linearity in deflection is obtained along both axes, a single value of voltage will not result in the same magnitude of spot displacement vertically and horizontally. This modification of the original presentation does not alter the *fundamental* facts which were being conveyed. They still remain as stated, except that we must recognize the difference in sensitivity or deflection-factor ratings of the two deflection systems in the cathode-ray tube.

CHAPTER 9

LINEAR TIME BASES

(Sweep Circuits)

The greatest amount of the discussion presented thus far concerned the action of the deflection fields upon the electron beam and the positioning of the spot. We progressed through the application of d-c voltages to the vertical and horizontal deflection plates and made brief mention of the application of a varying or a-c voltage to the deflection plates.

Now, the primary function of the cathode-ray tube as used in oscilloscopes, which are employed in connection with most types of electrical equipment, is the observation of a-c phenomena of one type or another. When we speak of electrical equipment, we include communication instruments, such as radio receivers, transmitters, audio amplifiers, television, and radar apparatus; also electrical test equipment, industrial control equipment, and many other similar types of apparatus, too numerous to mention here. As far as the oscilloscope is concerned, the first step that we have to consider in this a-c observation is the method of spreading out the a-c wave. This is necessary in order that the waveform of alternating voltages and currents can be examined by viewing the screen of the cathode-ray tube.

This observation requires that some means be available whereby the electron beam, which has been placed under the influence of the phenomenon to be examined or observed, will be caused to move in accordance with a predetermined velocity. In other words, the electron beam moves in a definite direction with a certain speed. Since speed is a function of distance and time, the electron beam will traverse a path across the screen of the tube in a certain amount of time. This path that the beam travels is called the *time base* and will permit the investigation of electrical waveforms because it will help determine the variation in amplitude of the waveform with respect to time. Thus, we find that a time base is necessary in order to give the pattern or image to be observed two dimensions, one time and the other amplitude. In order to help produce this time base, a special circuit, called the *sweep circuit,* has to be employed in conjunction with the cathode-ray tube.

The Time Base

In order to realize the importance of a time base in observing electrical phenomena, let us consider a simple sine wave. A sine wave, as well as all other alternating waves, contains three main electrical characteristics — namely, frequency, phase, and amplitude. At the moment, the quality that is of interest is the frequency of the signal. The alternating feature of the sine wave imparts a certain cyclic variation of the wave with respect to time, which variation we term the frequency of the signal. Thus when we say a signal has a frequency of 200 cps, it, of course, means that the wave is alternating in amplitude at the rate of 200 cycles each second. The time quality of frequency is not always stated when a frequency is mentioned, but is nevertheless understood to exist.

When we draw a sine wave graphically, we are actually showing a curve of amplitude versus time. In order to show this same sine wave on a cathode-ray tube, the same graphical relationship must exist, that is, the wave shown must also be a function of amplitude and time. It is the time base that helps produce this wave on the cathode-ray tube.

A question that might be raised at this point is the correlation between the time involved in the time base and that involved in the frequency of the actual wave. The time that it takes the electron beam to traverse a path across the screen is variable in all oscilloscopes. When we mention a 200-cps wave, we mean that in one second there are 200 cycles of alternation, in 0.5 second there are 100 cycles of alternation, and in 0.01 second there are 2 cycles of alternation. Thus, if a signal with a frequency of 200 cps is to be observed and the time base of the oscilloscope varied to the point where it takes the electron beam 0.02 of a second to complete its path from start to finish, only four complete cycles of the sine wave would appear on the screen. This relationship between the time base and frequency of the incoming signal will be more clearly evident as this chapter progresses.

The time base is formed by deflecting the spot across the screen from one point to another. This time-base deflection must position the spot at a certain point upon the screen, move it across the screen at a definite known rate, and instantly return it to its original position to begin a new cycle of time-base deflection. As mentioned before, it is the sweep circuit that helps deflect the electron beam to produce the time base. The exact length of this time base, as viewed alone on the screen, is a direct function of the sweep-circuit voltage and is sometimes referred to as the time baseline. The rate at which the spot moves across the screen is known as the *sweep frequency*. Application of the term "sweep" seems quite appropriate, since its action is to sweep the beam or spot across the screen.

From this brief analysis we see that the sweep circuit is a necessity in order to view waveforms on a cathode-ray tube. It is also known that the sweep circuit must be a signal-producing network of some sort because we have to deal with a sweep frequency which determines the time involved in moving the spot across the screen. By having a variable sweep frequency, any number of cycles of a particular waveform can be observed.

In most oscilloscopes, the signal to be observed on the cathode-ray tube is applied to the vertical deflection plates and the deflection voltage representing the time base is applied to the horizontal deflection plates. However, without any signal applied to the horizontal deflection plates (that is, with the time base absent), all that would be seen on the screen of the cathode-ray tube, due to the application of any shape waveform to the vertical deflection plates, would be a vertical line. If the signal input to the vertical plates is a sine wave, then the length of the vertical line observed would be equal to the peak-to-peak amplitude of the sine wave. We see a vertical line because the spot is moving up and down so rapidly, at a rate determined by the frequency of the signal, that the eye sees the movement as a single trace. If the frequency were low enough, the vertical line would appear to be flickering, indicating that the spot was not moving fast enough for the eye to see the movement as a single trace.

Fig. 9-1.—Time baseline generated by horizontal sweep signal.

When the sweep signal is applied to the horizontal deflection plates, there is also a horizontal force, the timing base, imparted to the electron beam, and hence the wave to be observed is spread out on the screen. In other words, the electron beam and hence the resultant spot will move in accordance with the amplitude of the wave and the instantaneous displacement with respect to time. If a sweep signal is applied to the horizontal deflection plates without any signal applied to the vertical deflection plates, then only a time baseline would appear on the screen of the cathode-ray tube. A photograph of such a time baseline is illustrated in Fig. 9-1.

The time base has been stressed in this section as being necessary in order to reproduce waveforms on the screen of the oscilloscope. However, the use of the oscilloscope is not limited to visual representation of the waveforms of certain signals, but has numerous other uses. In many of these applications, signals are applied to both the vertical and horizontal deflection plates, and the signal applied to the latter plates is not considered a time base in the same manner as previously discussed. This fact should always be remembered, because, in the remainder of this text, various other applications of the oscilloscope will be studied.

In order to completely understand how these various time bases are attained, it is necessary to study the sweep circuits and their associated signal waveforms that cause these time bases. It was stated that in order to reproduce a signal waveform on an oscilloscope, the signal to be observed is normally applied to the vertical deflection plates and the sweep signal to the horizontal deflection plates. The sweep signal can be represented by any voltage that will cause the desired result.

Simple D-C Sweep Circuit

The simplest means of illustrating the action of reproducing a waveform is to apply an a-c wave to the vertical plates and to apply a varying d-c voltage to the horizontal plates, which acts as the sweep signal. By arranging that the horizontal deflection takes place at a definite time rate, we establish a known time base. If this horizontal deflection were to be of a-c character, and if it were desired to establish a definite time base, a voltage of known frequency would be applied. However, that is for later discussion. In the meantime, let us sweep the beam across the horizontal axis, by means of a rotating potentiometer which is across a d-c supply, thereby developing the time base.

Let us examine Fig. 9-2, which is a basic illustration of the action of a d-c sweep. Part (A) represents a very low-frequency signal, for example, 1 cps, applied across the vertical-deflection plates. We select this frequency because it allows observation of the motion of the spot. Part (B) is a graphical illustration of the different spot positioning of the beam. The horizontal sweep voltage is developed across the potentiometer *P*. This potentiometer is arranged in the form of a circle with a rotating arm. You will note that contacts are provided and are numbered from *+14* to *0* to *−14*. The arm of the potentiometer rotates in a clockwise direction and you will have to visualize a minute jump which the arm makes as it rotates and moves from the *−14* to the *+14* terminal. You will have to assume the spacing between these two terminals to be extremely small, so that the

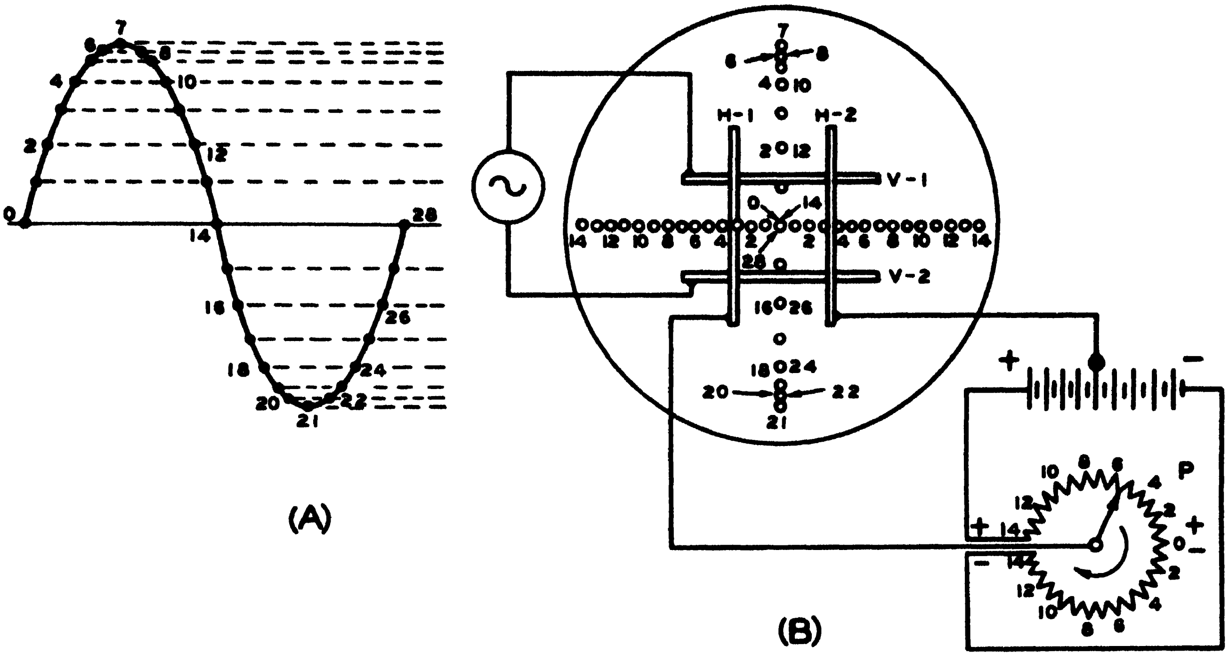


Fig. 9-2.—Graphical analysis of a sine-wave signal (A) on the vertical plates utilizing a potentiometer-driven horizontal sweep signal (B).

elapsed time interval during the jump is very small with respect to the time interval existing when the contact arm moves between the resistance steps or taps upon the potentiometer arm. You must also assume that this contact arm is rotated by a motor at a predetermined and constant speed.

The small individual circles, shown in part (B), indicate spot positions for the vertical and horizontal deflections of the beam when only *one* set of deflection plates is considered active at a time. In other words, the column of spot positions along the vertical axis exists without any horizontal deflection, and the row of spot positions along the horizontal axis exists without any vertical deflection. The reason for the uneven spacing of the circles along the vertical axis is that the amplitude variations of the signal applied to vertical plates are at a sinusoidal rate. The total length between the first and last vertical spot is equal to the peak-to-peak amplitude of the sine wave. The accuracy of the drawing indicates this. The points on the sine curve and those along the vertical axis on the screen are related, and if the dashed lines leading from these points of the sine curve were extended to the right, they would intersect the vertical axis circle of the same number.

Tracing Out First Half-Cycle

You will also note that while only seven vertical positions of the spot are shown above and below the horizontal axis, each spot, other than that which indicates the peak amplitude of the sine wave, appears twice during each half-cycle. This is indicated by each spot lining up with two points on the sine wave. To foster closer association between the vertical and horizontal deflections, when both exist, we have indicated the even-numbered vertical-deflection spot positions only. The odd-numbered positions were omitted for the sake of clarity, but they are, nevertheless, understood to exist. The numbers adjacent to the vertical column of spot positions correspond with the numbers shown upon the wave. In other words, the zero position, or normal center position of the spot, corresponds to numbers *0, 14,* and *28* on the wave. Spot position *2* on the rising quarter of the cycle is also spot *12* on the second quarter. Spot positions *4* and *10* coincide, because they represent equal displacement from the normal position, or horizontal axis. The same is true of spot positions *6* and *8*. Spot position *7* occurs just once during the half-cycle, because it is the peak amplitude, and the voltage causes this displacement only once during each first half-cycle. In the second half-cycle, spot positions *16* and *26, 18* and *24,* and *20* and *22* are seen to coincide. Spot position *21* occurs only once during this half-cycle, because it indicates maximum displacement, which occurs only once during the second half-cycle.

You will further note that the horizontal-deflection sweep-voltage potentiometer has 14 taps on each side of the zero position, between the zero point and the two positions of maximum voltage. Even-numbered positions are only indicated here also, the odd numbers being omitted from the drawing for clarity reasons. The position of the spot upon the screen for each numbered position of the potentiometer rotor arm is indicated by corresponding numbers. The spots not numbered indicate the odd positions of the potentiometer. The speed of the potentiometer rotor arm is such that one complete revolution is made in the time required for the signal at (A) to complete one cycle, so that a single cycle appears upon the screen. When this condition exists, it is said that the sweep frequency corresponds with the frequency of the signal being observed. The effects of different sweep or timing axis frequencies will be shown as we progress through this chapter. Again we want to remind you that the rotor arm on the potentiometer moves from what is indicated as *−14* to *+14* in an extremely short interval of time. Also remember that the number of cycles appearing on the screen of the cathode-ray tube is determined by the frequency relation between the signal being observed and the frequency of the sweep. In the case now under discussion, the sweep frequency is determined by the number of revolutions per second that the potentiometer arm undergoes. You will soon learn why these facts are brought to your attention.

Do not forget that the vertical or horizontal spot positions indicated in this drawing only exist when the proper signal is applied to either set of deflection plates. When both the signal to be observed and the sweep voltage are applied simultaneously, each to its own set of deflection plates, two forces are acting upon the electron beam, each force displacing the beam in its required direction. Hence, the spot, for any one instant of time, occupies a completely different position on the screen and, in the course of a small amount of time, traces out the waveform to be observed.

Image Formation

Let us now see how the image is developed upon the screen by actually combining the two individual deflections shown in Fig. 9-2B. We shall assume that the rotor arm of the potentiometer is initially set at the *+14* tap when the motor starts and that this position coincides wih the start of the a-c cycle of the *0* point on the sine wave. With the rotor arm at *+14,* horizontal deflection plate *H-1* is 14 units positive and in accordance with Fig. 9-2B, the beam spot at that instant has a horizontal displacement of 14 units, all the way to the left of *H-1* but has no vertical displacement; hence the spot occupies the position *0-14* indicated in Fig. 9-3.

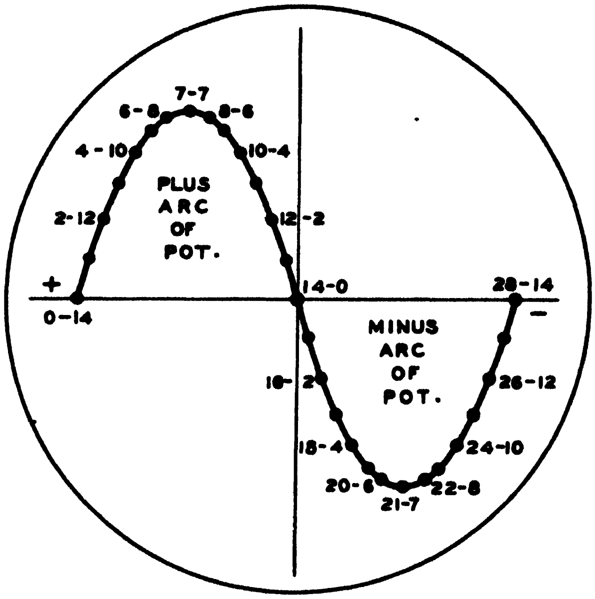


Fig. 9-3.—Graphical analysis of a sine wave correlating the instantaneous position of the potentiometer arm supplying the horizontal sweep signal with the instantaneous sine-wave voltage.

Next consider the instant when the rotor arm is at the *+12* tap and the amplitude of the sine wave has advanced to position *2.* The displacement of the spot is then 2 units up along the vertical axis and plus 12 units along the horizontal axis and the resultant spot position is shown in Fig. 9-3 as point *2-12.* The first of these numbers designates the vertical displacement due to the sine wave and the second number represents the horizontal displacement due to the potentiometer voltage, in each case with respect to the numbers indicated in Fig. 9-2. (Remember that the *same* amount of time has elapsed for the potentiometer arm to rotate from the *+14* tap to the *+12* tap as the sine wave of voltage to increase in amplitude from the *0* point to the number *2* point.)

We can illustrate the combination of the horizontal and vertical displacement of the spot very easily by the use of plane vectors. The vector diagram of Fig. 9-4 indicates the combination when the sine wave is at its number *2* position and the potentiometer is at its *+12* position. Vector *0-2* represents the vertical displacement due to the sine wave and its magnitude indicates the relative amplitude of the sine wave at the number *2* position of the wave. The vector *0-12* represents the horizontal displacement of the spot due to the potentiometer or sweep voltage and its magnitude indicates the relative amplitude of the sweep voltage when the potentiometer is at the *+12* position. If these two vectors are added, the resultant vector will show the position of the spot. In the diagram of Fig. 9-4, line *O-A*

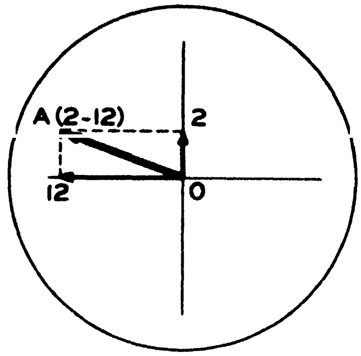


Fig. 9-4.—Illustration of the resultant spot displacement by combination of vertical and horizontal displacements (vectors).

is the resultant vector and point *A*, the extreme end of the vector, represents the position of the spot and is indicated as *2-12*.

Next let us consider the time when the potentiometer rotor arm is at *+10* tap and the amplitude of the wave has advanced to point *4*. The displacement is 10 units along the horizontal axis and 4 units along the vertical axis and the resultant position of the spot is *4-10* as indicated in Fig. 9-3. The potentiometer arm continues revolving, and when it makes contact with tap *8*, the amplitude of the sine wave has advanced to point *6;* together these two voltages cause a resultant displacement of the beam spot to the position *6-8*. At the peak of the sine wave, the arm of the potentiometer is midway between the *+8* and *+6* taps, or theoretically at the *+7* tap, although this number is not illustrated in the drawing. The resultant spot at this instant of time is in position *7-7* as seen in Fig. 9-3. After reaching its peak at point *7*, the amplitude of the first half-cycle of a-c voltage starts decreasing and numbered points *8, 10, 12,* and *14* are traced out. At the same instants of time, the potentiometer, which continues revolving in a clockwise direction, makes contact with numbered taps *6, 4, 2,* and *0* respectively. The resultant displacement of the spot due to the combination of these individual vertical and horizontal displacements is indicated by respective spot numbers *8-6, 10-4, 12-2,* and *14-0* in Fig. 9-3. At the moment when the voltage of the a-c signal reaches the point marked *14*, which is zero voltage, the potentiometer arm has also reached its zero voltage tap, so that there is no deflection force exerted upon the beam. Consequently, the spot remains in its normal center position, which on the drawing of Fig. 9-3 is represented by the *14-0* spot.

Tracing Out Second Half-Cycle

You now have seen how the first half-cycle of the wave is traced out on the screen of the cathode-ray tube. The upper half of the straight-line image that exists when the spot has only vertical displacement and no horizontal displacement has now been spread out with a time base or sweep. Let us now develop the remaining half-cycle of the sine wave. The lower half-cycle of the sine wave, which makes deflection plate *V-2* positive with respect to plate *V-1,* helps cause the spot to move downward, below the horizontal axis. At the same time, the movement of the potentiometer rotor arm, past the *0* point (still in the clockwise direction), causes deflection plate *H-2* to become positive with respect to plate *H-1*. The combined action of these two forces is such that the spot will always be below the horizontal axis and to the right of the vertical axis. As the potentiometer rotor arm moves to *−2*, the amplitude of the sine-wave voltage increases in the reverse (negative) direction from point *14* to point *16* and the resultant displacement of electron beam due to these two forces places the spot in position *16-2* as seen in Fig. 9-3. Further motion of the rotor in synchronism with the voltage wave causes resultant spot displacements as indicated by spot numbers *18-4, 20-6, 21-7, 22-8, 24-10, 26-12,* and *28-14*. The last position is zero voltage or deflection along the vertical axis, but maximum deflection, to the right, along the horizontal axis.

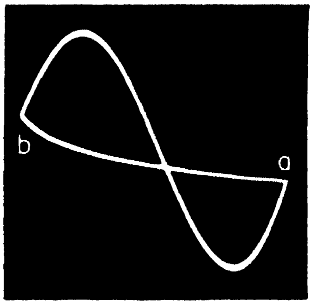


Fig. 9-5.—Actual photograph from the face of a cathode-ray tube of one cycle of a sine wave.

Continued operation of the rotor and application of the voltage wave repeats the process, for, as previously stated, an extremely short interval after the spot *28-14* has been positioned, the horizontal timing voltage sweeps back to its original position of *+14*, which is in time with the start of the cycle, at point *0*. The rate at which this return takes place is usually made to occupy a time interval which is from 1/25th to 1/100th of the time required for the completion of a single cycle.

In Fig. 9-5 there is shown an actual photograph of one cycle of a sine wave as reproduced on the screen of the cathode-ray tube. The sweep signal and input sine wave both have the same frequency. Although an electronic type of sweep circuit is used in the oscilloscope where this photo and others following are taken, for the sake of discussion, we will assume that the sweep voltage is secured from the potentiometer of Fig. 9-2. That part of the photo between points *a* and *b* represents the

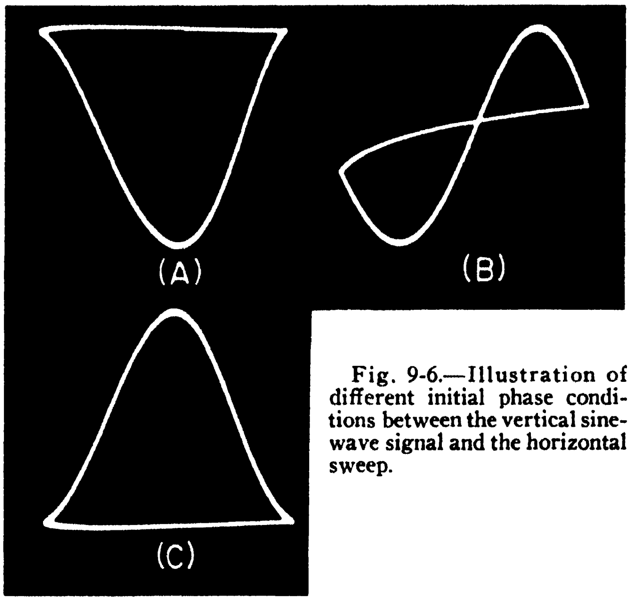


Fig. 9-6.—Illustration of different initial phase conditions between the vertical sine-wave signal and the horizontal sweep.

travel of the beam during the time the rotor arm travels from the *−14* to *+14* tap. It is known as the *retrace* line. In order to reproduce the picture as shown, the start of the sweep signal is made at the *+14* tap of the potentiometer at the same time the input sine wave starts at zero. When in this condition, the two voltages are said to be *in phase*, as mentioned previously. What happens to the reproduced picture when the starting-time relationship between the sweep voltage and input sine wave is no longer in phase?

Different Phase Conditions

In Fig. 9-6 there are shown three different pictures representing various phase conditions, or starting-time relationships, between the sweep and sine-wave voltage of Fig. 9-2. In each case, the frequency of both voltages is the same; that is, the time required to complete one revolution of the potentiometer rotor arm is the same as that required to complete one cycle of alternation of the sine wave. Photograph (A) of Fig. 9-6 can be representative of the case where the start of the sine wave of Fig. 9-2 is one-quarter of a cycle ahead of its zero point or at point *7*, and the start of the potentiometer rotor arm is at *+14*. This is the same as saying that the input sine wave is leading the sweep signal by 90°. Another way of looking at the reproduction of this picture is to assume that the rotor arm started moving from in between taps *−6* and *−8* (which point can be considered as *−7*) at the time the sine wave started at its zero point.

The wave at (B) of Fig. 9-6 will result if the sine wave of Fig. 9-2 started at point *14* and the rotor arm at tap *+14*, or if the rotor arm started moving at point *0* at the same time the sine wave also started at point *0*. In either case the same picture will be produced and the two voltages are said to be 180° out-of-phase. The reproduced picture at (C) of Fig. 9-6 will result if the sine wave of Fig. 9-2 started at point *21* and the rotor arm at the *+14* tap, or if the rotor arm started moving from in between *+8* and *+6* (considered as *+7* tap) at the same time the sine wave started at point *0*. Remember that, in each case, the sine wave must complete one cycle of alternation and the rotor arm one revolution from their starting points.

You have seen how the time base is introduced and how just a vertical straight-line image of amplitude, when the sine wave of voltage is applied to the vertical-deflection plates, becomes an image of the sine wave itself by simultaneously applying the horizontal time-base deflection. At this point, we want to mention that the time base shown here is not supplied commercially with the type of apparatus described. However, this description will serve to explain the manner in which the sweep voltage, developed by some other means, spreads the wave. Also it might be well if you understood that the slow speed of the rotor arm and the low frequency of the a-c wave were purposely selected in order to expedite comprehension. With both frequency and potentiometer rotor speed increased, it is impossible to follow the movement of the spot, because it moves so rapidly, but the positioning of the spot remains the same. The image, instead of being shown as a spot, which moves from place to place, assumes the form of a solid line, of the waveshape shown.

Variation In Sweep Frequency

An oscillogram in which just one cycle appears upon the screen is obtained, as mentioned previously, when the time or frequency of the sweep is the same as that of the phenomenon being observed. What happens if the time of the sweep frequency is twice as great as that of the wave being observed or if it is one-half the frequency of the wave being observed? Let us analyze the former where the frequency of the sweep circuit is twice the frequency of the voltage wave being examined. This means that the potentiometer rotor arm makes two revolutions during the period required to complete one cycle of the wave being observed. In the case being discussed, the potentiometer rotor arm completes two revolutions during one second.

During this discussion, we must assume that a certain phase condition exists between the varying d-c voltage applied to the horizontal-deflection plates and the a-c wave applied to the vertical-deflection plates. The time elapsed between the start of the potentiometer arm and the start of the a-c wave can be expressed

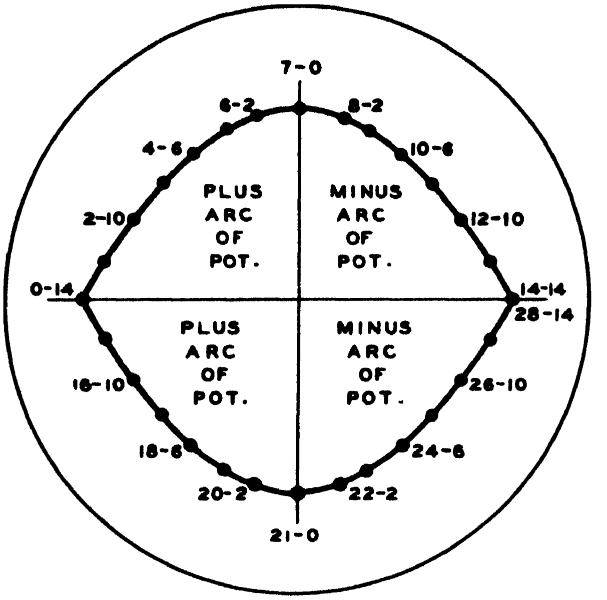


Fig. 9-7.—Graphical representation of the case where the sweep frequency is twice that of the sine-wave signal under observation.

in terms of a frequency or phase difference. In this analysis, we are assuming that the potentiometer rotor arm starts moving simultaneously with the start of the a-c wave and in this manner the two deflection forces are said to be in phase. With the wave and the arm moving at their constant frequencies, the two voltages will stay in phase.

The basic data are furnished in Fig. 9-2. For proper comprehension of what is being explained, it will be necessary to correlate Figs. 9-2 and 9-7. Since the potentiometer rotor moves twice as fast as the voltage wave under observation, it completes half of one revolution during the time that the voltage wave completes one-fourth of one cycle. Referring to Fig. 9-2, the rotor starts moving at the *+14* tap when the voltage wave is at zero. There is no vertical displacement, but only a horizontal displacement toward plate *H-1* because this plate is positive with respect to plate *H-2*. This condition places the spot at point *0-14* in Fig. 9-7.

By the time that the voltage wave has advanced to point *2* in Fig. 9-2, the rotor arm has advanced to tap *10;* this represents a vertical displacement of 2 units and a horizontal displacement of 10 units, resulting in the positioning of the spot at *2-10* in Fig. 9-7. By the time the wave has advanced in amplitude to point *4*, the rotor has moved to tap *6* and the resultant displacement is shown as point *4-6*. Maintaining the same rate of rotation, the rotor is at tap *2*, when the wave amplitude has reached point *6*, thus positioning the spot as *6-2*. When the wave reaches its maximum amplitude, point *7* in Fig. 9-2, the rotor arm has reached the zero tap. With no horizontal deflection and 7 units vertical deflection, the spot is positioned on the vertical axis, as *7-0* in Fig. 9-7. From this latter figure, it is evident that one quarter-cycle of the wave has been covered for half of a revolution of the rotor arm.

The amplitude of the first half-cycle of the wave now decreases and is seen to pass through the numbered points *8, 10, 12,* and finally to *14*, which is zero voltage. At the same time, the rotor arm is advancing in its arc of travel. Having passed its zero point, it is making horizontal-deflection plate *H-2* positive with respect to *H-1*, so that the spot is positioned to the right of the vertical axis reference line. Furthermore, the wave is still above the horizontal axis reference line and vertical-deflection plate *V-1* is still positive with respect to *V-2*, so that the spot positions are still above the horizontal axis. During the time that the wave is passing through its second quarter-cycle, the rotor arm is passing through the second half of its first complete revolution and new spot positions occur as seen by points *8-2, 10-6,* and *12-10,* and as the wave reaches zero voltage again, designated as *14* in Fig. 9-2, the rotor arm is at *−14;* the final spot reached is *14-14*, as seen in Fig. 9-7.

At this instant, the rotor arm moves from *−14* to *+14*, which means that the sweep goes back to its original starting point and makes horizontal-deflection plate *H-1* 14 units positive with respect to *H-2* and brings the spot back to the *0-14* position, as shown in Fig. 9-7. If you can view this return from the *−14* tap to the *+14* tap on the potentiometer rotor as taking place in an extremely small interval of time, it can be said that there is no time lost between the movement of the horizontal sweep from spot *14-14* in Fig. 9-7 to spot *0-14* in the same figure, with respect to the continued motion of the wave. (In actual practice, when using electronic sweeps, some time is always lost and this loss in time is evident in the oscillographic waveforms included herein.) According to the sine wave illustrated in Fig. 9-2, it now starts upon its half-cycle below the zero reference line. When so doing, it makes vertical-deflection plate *V-2* positive with respect to *V-1*, so that the beam is deflected toward the former plate and thus, in conjunction with the horizontal voltage, the spot is positioned below the horizontal axis. At the same time, the rotor arm, moving on the plus side, also positions the spot to the left of the vertical axis. By the time the wave has increased in amplitude (in the negative direction) to point *16* in Fig. 9-2, the rotor arm has moved to tap *+10* and the spot is positioned as *16-10* in Fig. 9-7. During this third quarter-cycle of the wave, the rotor arm moves over the entire plus half of the potentiometer, from *+14* to *0*, and the spot is seen to be positioned as *18-6*, *20-2*, and *21-0*. When the maximum amplitude is reached during the second alternation of the wave, the rotor arm has reached zero. Since there is no horizontal displacement and there is maximum vertical displacement at this latter point, the spot is positioned on the vertical axis. The wave now starts on its last quarter-cycle while the rotor arm starts advancing from *0* to the *−14* tap. The sweep voltage now makes horizontal-deflection plate *H-2* positive with respect to *H-1*, so that the spot, for the remainder of the sine wave, is positioned below the horizontal axis and to the right of the vertical axis. The resultant displacement causes the spot to move through positions *22-2*, *24-6*, and *26-10*, and when the wave reaches its zero voltage point, number *28* in Fig. 9-2, the rotor arm is at *−14* and thus the maximum horizontal displacement with zero vertical displacement positions the spot at *28-14* as seen in Fig. 9-7. Continued operation of both the wave and rotor arm at these same frequencies repeats the cycle, and the recurrent action results in the repetition of the image shown.

Operation of the sweep at a frequency which is twice that of the input signal to be observed, when both voltages start at the time indicated, makes the reproduced picture symmetrical about either axis, as seen from Fig. 9-7. In other words, those parts of the picture above and below the horizontal axis, and also to the right and left of the vertical axis, are mirror images of each other. In Fig. 9-8A is an actual photograph of the pattern on the screen of a cathode-ray tube when the sweep signal applied to horizontal plates is twice the frequency of the sine wave applied to the vertical plates. Compare this photograph with the line drawing of Fig. 9-7.

The illustrations show what the reproduced picture of an input sine wave will appear like when the start of the sweep signal in Fig. 9-2 is made at the *+14* tap of the potentiometer at the same time the sine wave starts at zero. The photograph at (B) of Fig. 9-8 also represents a situation where the sweep frequency is equal to twice that of the input sine wave. In this case, however, the starting time for one of the voltages has changed. If the sine wave of Fig. 9-2 starts at point *7* instead of point *0*, and the rotor arm still starts from tap *+14*, the picture of Fig. 9-8B will result. The same picture will be reproduced if the rotor arm starts at tap *0* and the sine wave at its point *0*.

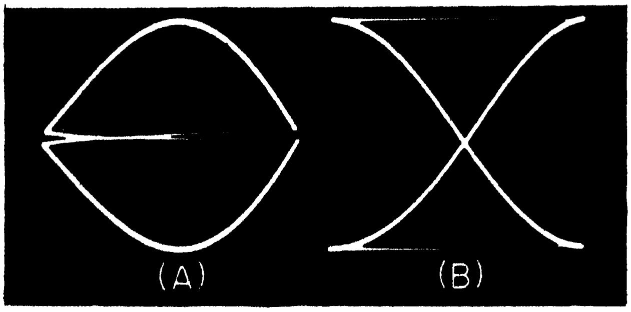


Fig. 9-8.—Actual photograph of a sine wave where the sweep frequency is twice that of the signal frequency. In (A) both signals started from their respective zero positions, while in (B) the sine wave started at its first voltage maximum, i.e., 90° ahead in phase from the sweep signal.

Other Frequency Relationships Between Sweep and Sine Wave

In the previous analysis, we illustrated what the reproduced signal looks like when the frequency of the sweep was equal to that of the sine wave and also when the sweep frequency was twice that of the sine wave. Let us now consider the case where the frequency of the sine-wave input is greater than that of the sweep signal. We will illustrate the case where the sine wave has a frequency equal to twice that of the sweep. In this case, the horizontal displacement of the spot occurs at a rate or speed which is one-half that of the sine wave or vertical displacement. This is the same thing as saying that the vertical displacement of the spot is twice as fast as the horizontal displacement.

Sine-Wave Frequency Twice Sweep Frequency

Again we refer to Fig. 9-2 as the source of the basic facts. The drawing of Fig. 9-9 is the illustration of the wave reproduced under the above-mentioned frequency conditions of the two voltages involved. In this analysis, the start of the sine wave is at *0* and that of the rotor arm at *+14* tap. Now, the wave under observation moves twice as fast as the sweep, which means that a complete cycle will occur during the time that the rotor arm is passing through one-half of a revolution. The result is the presence of two cycles upon the screen, as seen in Fig. 9-9. In this connection, we want to stress that the number of cycles which appear upon the viewing screen is a function of the frequency relation between the sweep frequency and the frequency of the wave being observed.

Since the wave is moving twice as fast as the sweep, the first half-cycle of the wave will be spread or swept by the first quarter of a revolution of the rotor arm, or one-quarter of the total sweep. This is evident in Fig. 9-9, where the wave moves through its half-cycle from *0* to *14*, during the time that the sweep moves from *+14* to *+7*. The second half-cycle of the sine wave is swept by the horizontal-deflection voltage as the rotor moves from this latter tap to tap *0*, so that a complete cycle of

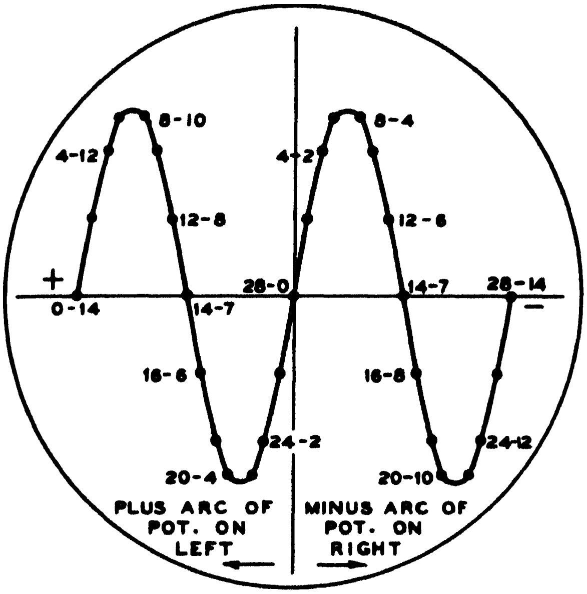


Fig. 9-9.—The signal (sine-wave) frequency is twice that of the sweep, but both signals start from their respective zero points.

sine wave is swept during half of the sweep cycle (that is, half a revolution of the rotor arm). As the rotor arm completes its revolution from the *0* tap to the *−14* tap, the sine wave starts all over again from points *0* to *28*. This results in the reproduction of the second cycle of sine wave seen in Fig. 9-9. The placement of the first cycle of sine wave is to the left of the vertical axis because during this period of time, horizontal-deflection plate *H-1* is positive with respect to *H-2*. The second cycle appears to the right of the vertical axis because during this period *H-2* is positive with respect to *H-1*.

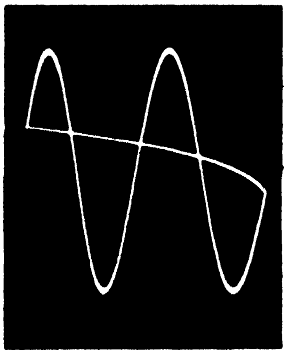


Fig. 9-10.—Actual photograph of the case analyzed in Fig. 9-9, where the sine-wave signal had twice the frequency of the horizontal sweep signal.

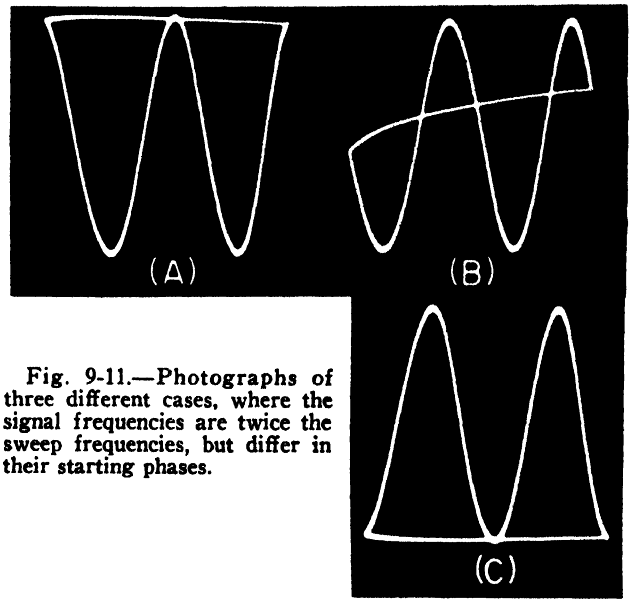


Fig. 9-11.—Photographs of three different cases, where the signal frequencies are twice the sweep frequencies, but differ in their starting phases.

A reproduced oscillogram of the conditions represented by Fig. 9-9 is shown in Fig. 9-10. Different patterns will be observed if the starting point of one of the voltages changes. The three photographs of Fig. 9-11 illustrate such a case. In each instance, the frequency of the input sine wave to the vertical-deflection plates is twice that of the sweep voltage. Let us consider the potentiometer rotor arm starting at the *+14* tap each time. Therefore, if the input sine wave is considered as starting at point *7*, then the picture reproduced will be that shown in part (A) of Fig. 9-11. If the sine wave starts at point *14*, then the picture shown at part (B) will result; the picture at part (C) appears when the sine wave starts at point *21*. These same pictures can be reproduced with the sine wave starting at zero for each case and the rotor arm starting at a different tap for each picture. These oscillograms, as well as those of Figs. 9-6 and 9-8B, can also be reproduced under different starting conditions than those mentioned.

Sweep Frequency Three Times Sine-Wave Frequency

Odd types of pictures result when the frequency of the sweep is even greater than twice that of the sine wave. In Fig. 9-12 are shown four different pictures that can result when the frequency of the sweep is equal to three times that of the input sine wave. Assuming that the sine wave of Fig. 9-2 starts at zero for each case, then in order to produce the picture at part (A) the rotor arm would have to start at the *+14* tap, for part (B) the rotor arm would start at the *+7* tap, for part (C) the rotor arm would start at the *0* tap, and for part (D) the rotor arm would start at the *−7* tap.

You will note that the photograph of Fig. 9-12A has letters next to certain parts of the waveform. These letters were inserted in order to illustrate the manner in which this waveform is produced. Remember that the frequency of the sweep is three times that of the input sine wave and that the sine wave starts at point *0* and the potentiometer rotor arm at the *+14* tap (see Fig. 9-2). When the rotor arm makes one complete revolution from *+14* to *−14*, only one third of the complete sine wave is traced out; this is indicated by curve

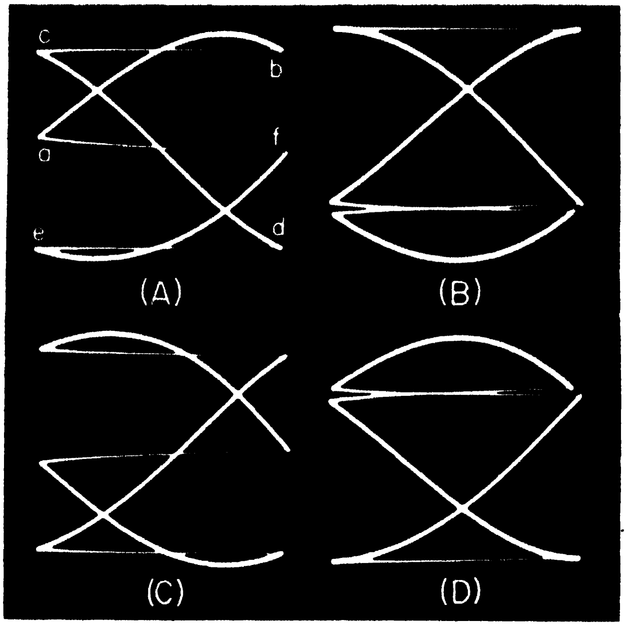


Fig. 9-12.—The frequency of the sweep signal is three times that of the input sine wave. The four oscillograms were obtained by starting the initial position of the potentiometer arm (horizontal sweep) at varying points with respect to the sine-wave signal.

a to *b* in Fig. 9-12A. In the short period of time when the rotor arm moves from the *—14* tap to the *+14* tap, the spot travels very rapidly from *b* to *c*.

Retrace

This latter period of movement is termed the *retrace* of the sweep signal. Some oscilloscopes may have provision for driving the grid of the cathode-ray tube beyond cutoff during this period of retrace time; this causes the electron beam, and hence the spot, to disappear for that time. There are, however, many oscilloscopes that do not use blanking during the sweep retrace time and in these units the spot appears as a line during sweep retrace. The intensity of this line depends upon the speed of retrace.

Now when the rotor arm starts on its second complete revolution going again from *+14* to *—14*, the middle third of the sine wave is traced out. This portion of the sine wave decreases in amplitude somewhat linearly from approximately point *8* to *20*, as seen in Fig. 9-2A. This over-all action causes the spot to trace out the curve from *c* to *d* of Fig. 9-12A. When the spot reaches *d*, the rotor arm has reached the *—14* tap and during the time the arm moves from *—14* to *+14* (the sweep retrace), the spot moves very rapidly from *d* to *e*. During the third complete revolution of the rotor arm, the sine wave completes its cycle and the spot travels from *e* to *f*. That portion of the photo from *f* to *a* again illustrates the retrace time.

Sweep Frequency Four Times Sine-Wave Frequency

The two reproduced photographs illustrated in Fig. 9-13 indicate two possible pictures when the sweep frequency is equal to four times that of an incoming sine wave. This means that for the d-c sweep of Fig. 9-2, the rotor arm must undergo four complete revolutions during the time the sine wave completes one cycle. This is the same as saying that for each complete revolution of the rotor arm, the sine wave traces out one-quarter cycle.

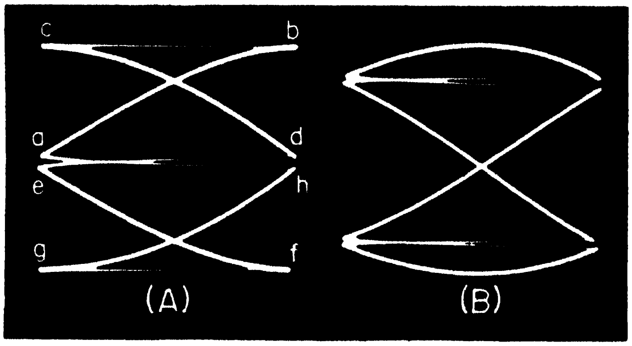


Fig. 9-13.—Two cases where the sweep frequency is equal to four times that of the sine-wave signal are illustrated in the oscillograms. In (A) both signals started at their respective zero points, while in (B) the potentiometer had already completed one-half rotation before the sine-wave signal was applied to the vertical plates.

The picture at (A) of Fig. 9-13 indicates the situation where the sine wave starts out at point *0* at the same time the rotor arm starts out at the *+14* tap. During the first complete revolution of the rotor arm, the sine wave moves from point *0* to *7* and the spot on the screen of the cathode-ray tube traces out line *a-b*. During the retrace time of the sweep, the spot travels from *b* to *c*. During the next revolution of the sweep from the *+14* tap to the *—14* tap, the sine wave traces out point *7* to point *14*. This over-all action causes the spot to trace out line *c-d* in Fig. 9-13A. During the next sweep retrace, the spot rapidly moves from *d* to *e*. For the next revolution of the rotor arm, the sine wave travels from point *14* to *21* and the spot traces out line *e-f*. The retrace then causes the spot to move quickly to *g*. During the last revolution of the rotor arm, the sine wave finishes its cycle from point *21* to *28* and the spot traces out line *g-h* as seen in Fig. 9-13A. When the rotor arm goes from the *—14* to the *+14* tap, the spot retraces from *h* to *a* and we are now back to the original starting position.

The photograph of Fig. 9-13B indicates the waveform that will be observed when, with the sine wave still starting at zero, the rotor arm starts at *0* instead of the *+14* tap.

At this time we again repeat that although the battery-potentiometer form of time base is not commonly used in commercial oscilloscopes, it is here employed as a means of establishing the foundation for comprehension of the means whereby the modern sweep circuits of different design provide the time base or horizontal deflection. The figures, which are shown as being developed when the sweep frequency is equal to half and other submultiples of the frequency of the wave under observation, are applicable to all forms of sweeps of substantially *linear* character. In other words, the sweep voltage should increase and/or decrease linearly with respect to time. The same is true of the pattern developed when the sweep frequency is some multiple of the frequency of the wave being observed. This should be borne in mind for, as you shall see later, modern sweep circuits are of the electron-tube variety.

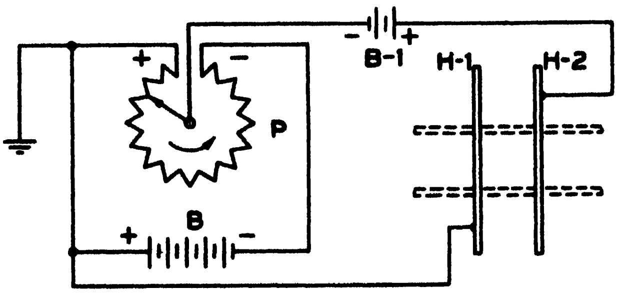
Perhaps it is premature to mention it at this time, but departure from linearity in the sweep circuit causes distortion of the image upon the screen. In other words, it is not a true reproduction of the actual wave applied to the deflection plates for investigation. More will be said about this subject later in the chapter.

Simple Mechanical Sweep

Before advancing to the actual electronic means of developing the sweep voltage or time base, it might be well to consider another form of a battery-operated sweep voltage. The sweep unit to be discussed can be considered as the equivalent of those mechanical sweeps that were used in some of the first types of oscilloscopes and is different from that previously described. Mechanical systems are also used to some extent in modern equipment requiring very low frequency sweep, such as certain types of radar and demonstration units. In the arrangement described, the normal position of the spot is off-center and it is located at one end of the tube screen along the horizontal axis.

A potentiometer *P* with a continually rotating contact arm is connected across the battery *B* of Fig. 9-14A. The arm can be considered as being connected to the shaft of some variable-speed motor so that the speed of rotation and hence the frequency of sweep can be varied. The design of this potentiometer is such that the arm jumps from the minus to the plus terminal in an extremely short interval of time (this is the retrace time) with respect to the time elapsed during its rotation around the resistor portion. A photograph of a specially designed potentiometer of this type appears in Fig. 9-14B. A separate supplementary spot-positioning battery *B-1* is connected as shown, with the plus terminal joined to horizontal deflection plate *H-2*. With the polarity of the main battery *B* connected as shown in Fig. 9-14, and with the arm of the potentiometer set at the positive end as the starting position, the action is as follows:

The negative terminal of battery *B-1* is grounded and since plate *H-1* is also grounded, this effectively puts battery *B-1* across the horizontal-deflection plates. With the positive terminal of *B-1* connected to *H-2* the beam spot is deflected toward plate *H-2* because this plate is positive with respect to plate *H-1*. This positions the spot on the horizontal axis, but quite distant from the exact center of the screen. As the arm starts rotating in the direction indicated, the voltage from battery *B* starts *bucking* the voltage of battery *B-1*, and the potential difference existing between *H-2* and *H-1*, the former positive with respect to the latter, becomes less and less; hence the spot moves toward the center of the screen away from plate *H-2*. When the rotor arm is about midway between the ends of the potentiometer, the bucking voltage offered by battery *B* can be considered as being equal to the voltage of *B-1* and, at this instant, the spot is about in the center of the screen. As the rotor arm continues to move toward the negative end of the potentiometer, the voltage offered by battery *B* is such that it more than bucks out the voltage of battery *B-1*, which makes deflection plate *H-1* positive with respect to *H-2*, and the spot moves toward *H-1*.



(A)

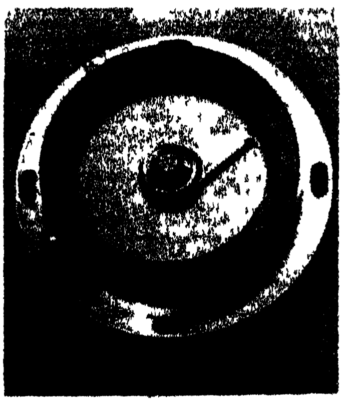


Fig. 9-14.—(A) shows the connections of a simple mechanical potentiometer drive that supplies a continuous horizontal sweep voltage. (B) is a photograph of the specially designed potentiometer for generating a horizontal sweep voltage.

At the instant when the rotor arm reaches the most negative point on the potentiometer resistance strip, *H-1* is most positive with respect to *H-2;* the beam spot has moved the greatest distance from its original position and has swept across the screen. At the next instant, an extremely small interval of time, as stated before, the arm jumps from the most negative point on the potentiometer strip to the most positive point on the strip and instantaneously returns the spot to its original starting position. If the speed of rotation of the arm is sufficiently great, and the rotation is recurrent, the trace of the beam spot constitutes a single line across the screen. The return sweep which occurs at an extremely rapid rate is barely visible.

Sawtooth Sweep

The production of time bases that are used for cathode-ray tubes in oscilloscopes are, in nearly all cases, completely electronic in operation. As mentioned previously, there are many different types of time bases that exist today, but the one most commonly used in oscilloscopes is the linear time base. There are numerous types of sweep circuits that can produce a linear time base, and in this chapter we will study the majority of linear sweep circuits that are employed in oscilloscopes.

With electronic sweep circuits, a wider range of sweep frequencies is possible than with the mechanical type; also the electronic type is much more versatile. Many other features, such as lower cost of construction, smaller amount of space occupied, and ease of operation, also indicate the shift toward electronic sweep circuits. In order to produce the electronic sweep voltage, some form of oscillatory circuit is invariably used. There are many different types of electronic circuits that are used to produce the sweep voltage. When dealing with electronic sweep circuits, it is the shape of sweep-voltage waveform which is the important characteristic. For most oscilloscopes, this sweep-voltage waveform must have a voltage-versus-time characteristic, that is, a single straight-line quality, which is linear, so that the beam spot travels across the screen at a constant speed.

At this point we would want to know something about the waveform of the linear sweep circuit. In order to produce an electronic sweep which has a linear time base, there is essentially only one type of linear sweep voltage that meets all requirements today. A fair idea of the shape of this sweep voltage waveform can be obtained from studying the simple mechanical sweep circuit of Fig. 9-14A from a different angle.

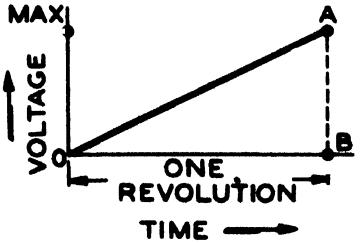


Fig. 9-15.—Graphical representation of the voltage-time curve produced by the mechanical potentiometer sweep, shown in Fig. 9-14.

We already know that this circuit will produce the desired linear time base. Therefore, if we can graphically illustrate the voltage-versus-time characteristic of the sweep signal applied across the horizontal deflection plates, we will have a good idea of the shape of the sweep voltage required by most linear electronic sweep circuits. Let us assume that the polarities of the batteries in Fig. 9-14A are reversed. This means that as the rotor arm of the potentiometer revolves in its counterclockwise direction, the voltage applied to the deflection plates, due to battery *B,* will be increasing. This increase in voltage is at a constant rate because of the constant velocity of the rotor arm and is, therefore, linear. A graphical plot of the voltage-versus-time relationship is illustrated in Fig. 9-15. At the starting point of the rotor arm, the voltage applied, due to battery *B,* is zero. As the arm revolves, the voltage increases at a constant rate until it reaches a maximum at the end of one revolution of the rotor arm. This maximum voltage is shown as point *A* in the drawing. When the rotor arm completes its revolution, an extremely short interval of time elapses for the arm to jump from one polarity terminal to the other. This is the same as saying that the voltage, after reaching its maximum point, collapses to a zero value in this short interval of time. This collapse of voltage is illustrated by the dashed line from points *A* to *B*. This line is drawn perpendicularly, indicating the collapsed time to be instantaneous, which is only the *ideal* situation. The over-all shape of this sweep voltage, which is in the form of a sawtooth wave, is that required from the linear electronic sweep circuits that are used today.

There are quite a few electronic sweep circuits that are designed to produce sawtooth waves for use in oscilloscopes. These circuits fall into three main categories — those of the gaseous (thyratron) relaxation-oscillator type, those of the multivibrator type, and finally those of the blocking-oscillator type. Of these three classifications, the thyratron and multivibrator are used most often in commercial cathode-ray oscilloscopes.

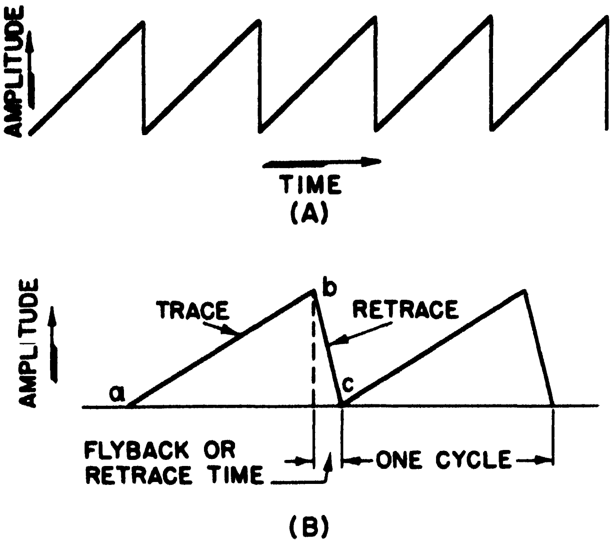


Fig. 9-16.—(A) shows an ideal sawtooth pattern having zero retrace time, while (B) shows the drawing of an actual (nonideal) sawtooth signal.

A drawing of an ideal sawtooth signal, showing five cycles of the wave, is illustrated in Fig. 9-16A. The heavy diagonal lines indicate the rising voltage characteristic which sweeps the beam across the screen. The perpendicular lines, which are drawn lighter, indicate the *retrace time* of the beam spot from its final position at one end of the cathode-ray-tube screen back to its starting position. The retrace time is also commonly known as the *flyback time*. The retrace time, or flyback time, must occupy a small portion of each cycle in order for the sawtooth to be useful as the sweep. If the retrace occupies a considerable portion of the time of each cycle, the complete waveform under observation will not be reproduced. In the photographs of Figs. 9-5 and 9-11B, the time of the sawtooth retrace is appreciable enough to prevent the complete sine wave from being reproduced. What actually happens is that the maximum sweep voltage has been reached before the input sinusoidal signal has traced itself out. However, the retrace time in these photos is not long enough to be objectionable.

The ideal sawtooth wave is a complex signal made up of numerous harmonics. It contains all even and odd harmonics of sine terms only.[1] The amplitudes of the various harmonics decrease in inverse proportion to the number of their harmonic. In other words, the amplitude of the second harmonic is ½ that of the fundamental, that of the third harmonic is ⅓ the amplitude of the fundamental, and so on. All the even harmonics contain negative sine waves, that is, their waves are shifted 180° with respect to those of the fundamental and all odd harmonics.

In the sawtooth wave where the retrace time is not zero (as in the actual case), but is some percentage of the complete wave, the amplitudes of the harmonics do not decrease inversely as the order of harmonic. The manner of amplitude decrease depends on the percentage of retrace. For retrace times of ten or twenty per cent of the nonideal sawtooth period, the amplitudes of the first four or five harmonics are, in general, comparable to those of the ideal sawtooth. However, the amplitudes of the higher-order harmonics decrease much more rapidly than those of the ideal wave.

A line drawing of two cycles of a typical sawtooth wave appears in Fig. 9-16B. The linear part of the wave, rising from *a* to *b*, often called the ***trace***, sweeps the spot across the screen; the part from *b* to *c* is the *retrace*, which returns the spot to its original starting position. The time for the return of the spot is marked off on the drawing and is indicated as the flyback or retrace time.

If the sawtooth signal takes on the shape similar to the drawing of Fig. 9-16A, then the sweep signal is said to have very good linearity and a fast retrace time. Such type sawtooth waves are always desired because they help reproduce the waveform to be observed as best a duplicate of the input signal as possible. Any defects in the sawtooth wave will show up in the reproduced picture.

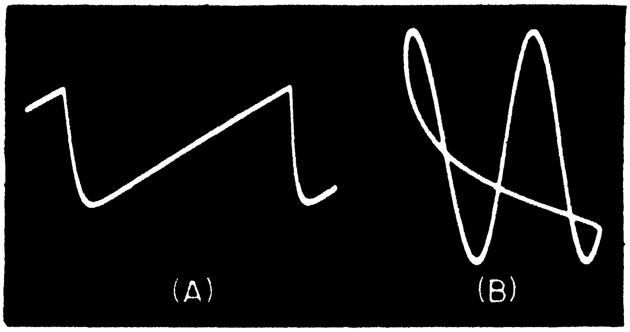


Fig. 9-17.—The imperfect sawtooth wave, having a long retrace time, shown in (A) produces the incompleted sine-wave pattern shown in (B).

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[1] The equation of an ideal sawtooth wave indicating the harmonic content is equal to

$$\frac{E}{\pi}\left(\sin x - \tfrac{1}{2}\sin 2x + \tfrac{1}{3}\sin 3x - \tfrac{1}{4}\sin 4x + \dots + (-1)^{n-1}\frac{1}{n}\sin nx\right)$$

where E equals the peak-to-peak amplitude of the signal and n the order of harmonic.

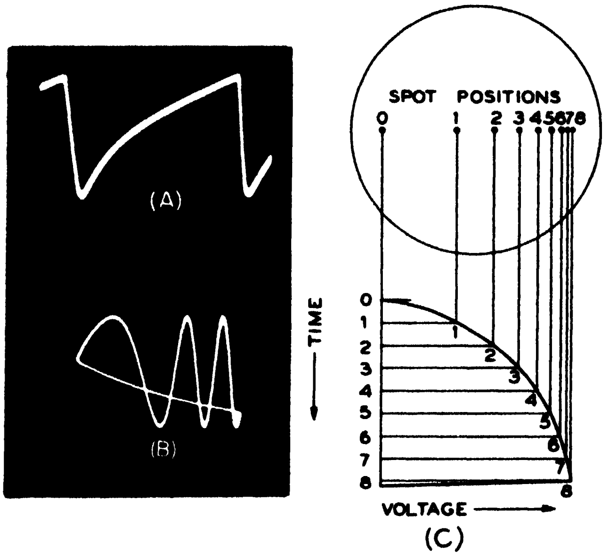

Fig. 9-18.—The distorted sine-wave pattern shown in (B) is produced by the nonlinear sawtooth signal, illustrated by the oscillogram in (A); (C) is a graphical representation of spot travel for a nonlinear sweep signal.

Defects in the retrace time and in the trace of the sawtooth are equally as common. If the retrace time occupies a high percentage of the time required to complete one cycle of sawtooth, as mentioned previously, parts of the input wave will not be reproduced. This is further illustrated by the two pictures of Fig. 9-17. That at (A) is a photograph of the actual sawtooth sweep used to reproduce two cycles of a sine wave. The reproduced picture is shown by the photograph at (B). Because too much time elapsed in the retrace of the sawtooth at (A), the sine wave at (B) appears incompleted.

If the rising part of the sawtooth, that part reproducing the input signal, is not linear, then the signal to be observed will not be an exact reproduction of the input. This is best illustrated by actual photographs as shown in Fig. 9-18. The photo at (A) shows a nonlinear sawtooth sweep signal and the photo at (B) is a sine wave reproduced by the sweep at (A). Note how the nonlinearity of the sweep has caused the cycles of

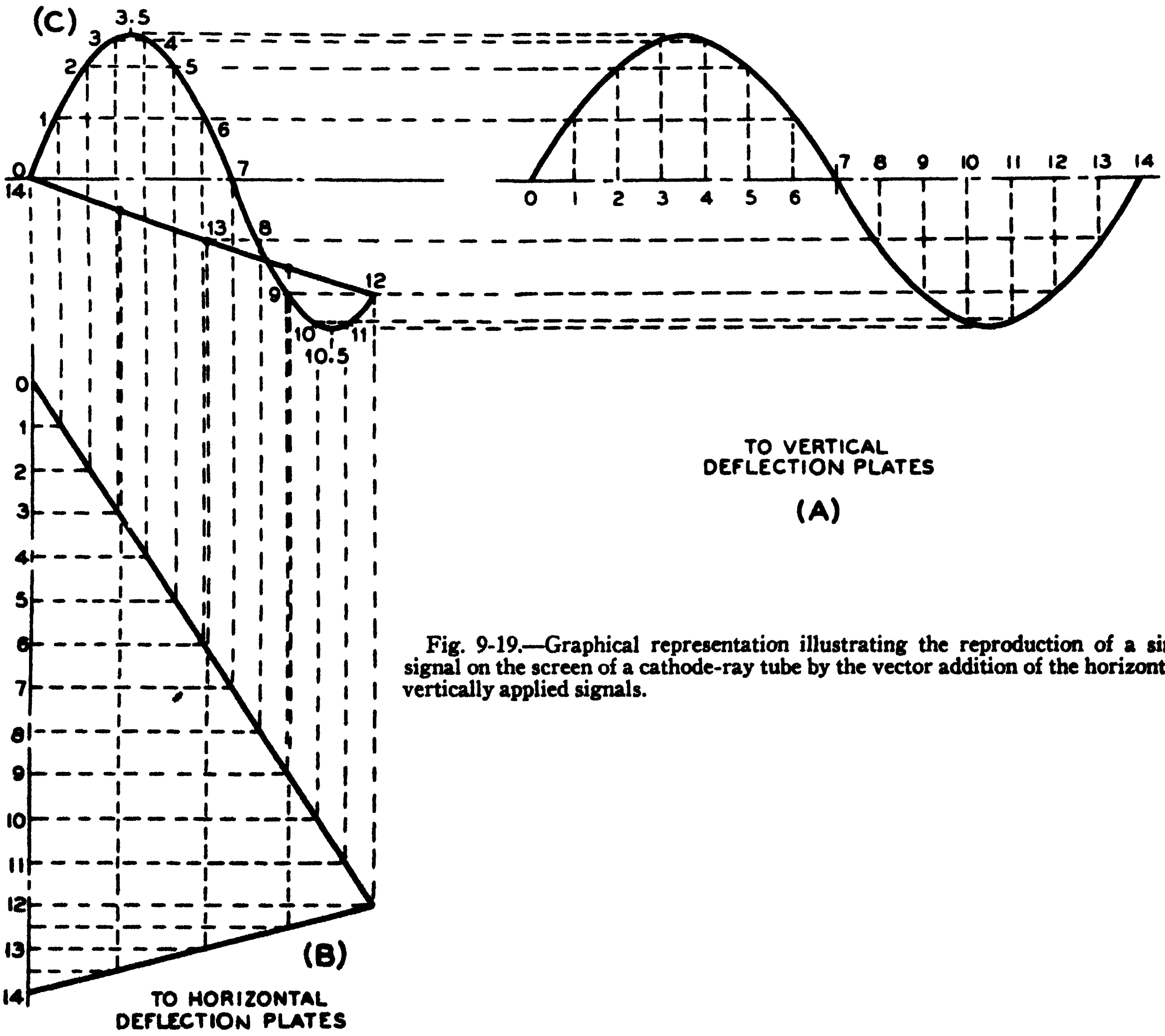

Fig. 9-19.—Graphical representation illustrating the reproduction of a sine-wave signal on the screen of a cathode-ray tube by the vector addition of the horizontally and vertically applied signals.

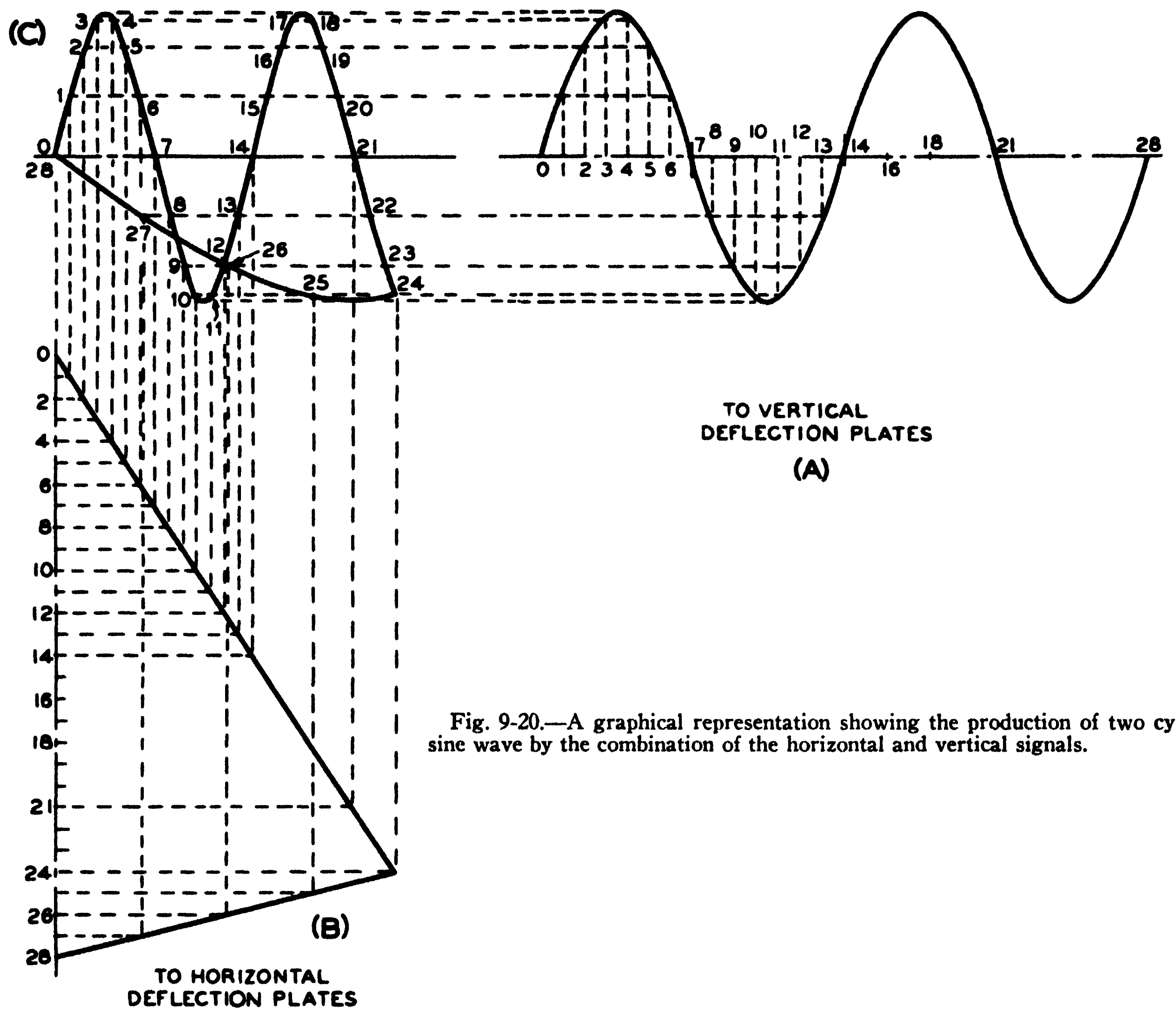

Fig. 9-20.—A graphical representation showing the production of two cycles of a sine wave by the combination of the horizontal and vertical signals.

sine wave to become unevenly spaced along the time axis. The graphical illustration in Fig. 9-18C is a basic method showing how the spot travels across the screen during a nonlinear sweep. The single cycle of sawtooth sweep has the nonlinearity of its curve exaggerated to better portray the case under discussion. We are assuming that the frequency of sawtooth is constant and the divisions along the time axis of the sweep signal are all equal to each other.

At point *0*, the very beginning of the sweep, the spot is on the extreme left of the screen. As the sweep advances, the spot will move from left to right. If we assume that the sweep is at a very low frequency and that it is only effective every ⅛ of the charging time, then we can visualize individual spot positions during each time displacement. For the first ⅛ of the cycle, the amplitude of the sweep has its greatest increase and the spot travels from point *0* to *1*. For the next ⅛ of a cycle, the amplitude of the sweep, between points *1* and *2*, has decreased, and hence the spot has traveled a shorter distance. For each succeeding ⅛ of a cycle, the change in amplitude from the previous point to the next one is constantly decreasing. Consequently, the spots positioned at the right side of the screen will be bunched and those at the left side spread out. The spreading and bunching of the spots is, however, gradual, and there is no specific line of demarcation.

When a sine-wave input is to be reproduced by such a sweep, the appearance of the picture will be similar to that of Fig. 9-18B. It should be remembered that it is the rate of amplitude increase of the sawtooth signal that determines the speed of the spot. Therefore, the picture of the unequal spot displacement does not depict unequal time variations as many would believe.

Before we analyze the various sawtooth-producing electronic sweep circuits, we will study various sawtooth waveforms and see how they function in reproducing sine-wave signal inputs.

Graphical Development of Sine Waves

In Figs. 9-19, 9-20, and 9-21 are illustrated graphical methods of producing one, two, and three cycles, respectively, of a sine wave upon the screen of a cathode-ray tube using a sawtooth sweep. In each drawing, part (A) represents the input sine wave to the vertical-deflection plates, part (B) the *single* cycle of sawtooth sweep applied to the horizontal-deflection plates, and part (C) the pattern that the spot traces out on the screen. The pattern caused by the retrace or flyback time of the sawtooth is included.

In these drawings, the voltages are divided into uniform time intervals in order to correlate the reproduced pattern with the applied deflection voltages in a similar manner to that of Fig. 9-2. The complete length of the time axis of the sine wave at (A), regardless of the number of cycles shown, is the same as that of the sawtooth at (B) in all three drawings. In each drawing, the time axis of both deflection voltages is divided into segments of the same amount to allow easier inspection of how the spot traces out the curve.

In the graphic reproduction of the curves in part (C), the following procedure is utilized:

Perpendicular lines, shown dashed in the drawings, are extended from the time axis of the sine waves at (A) and the sawtooth waves at (B) until each line intersects its respective curve. From these points on the individual curves, lines are then drawn at right angles from the original perpendicular lines. The direction of right angle is such that the newly extended lines from each wave will intersect each other. The intersecting points of the lines of the same number represent points on the reproduced curve. If enough of these lines are drawn, there will be sufficient intersecting points to accurately establish the appearance of the reproduced curve. Joining these intersecting points with a continuous line will reproduce the curves as shown at (C) in the drawings.

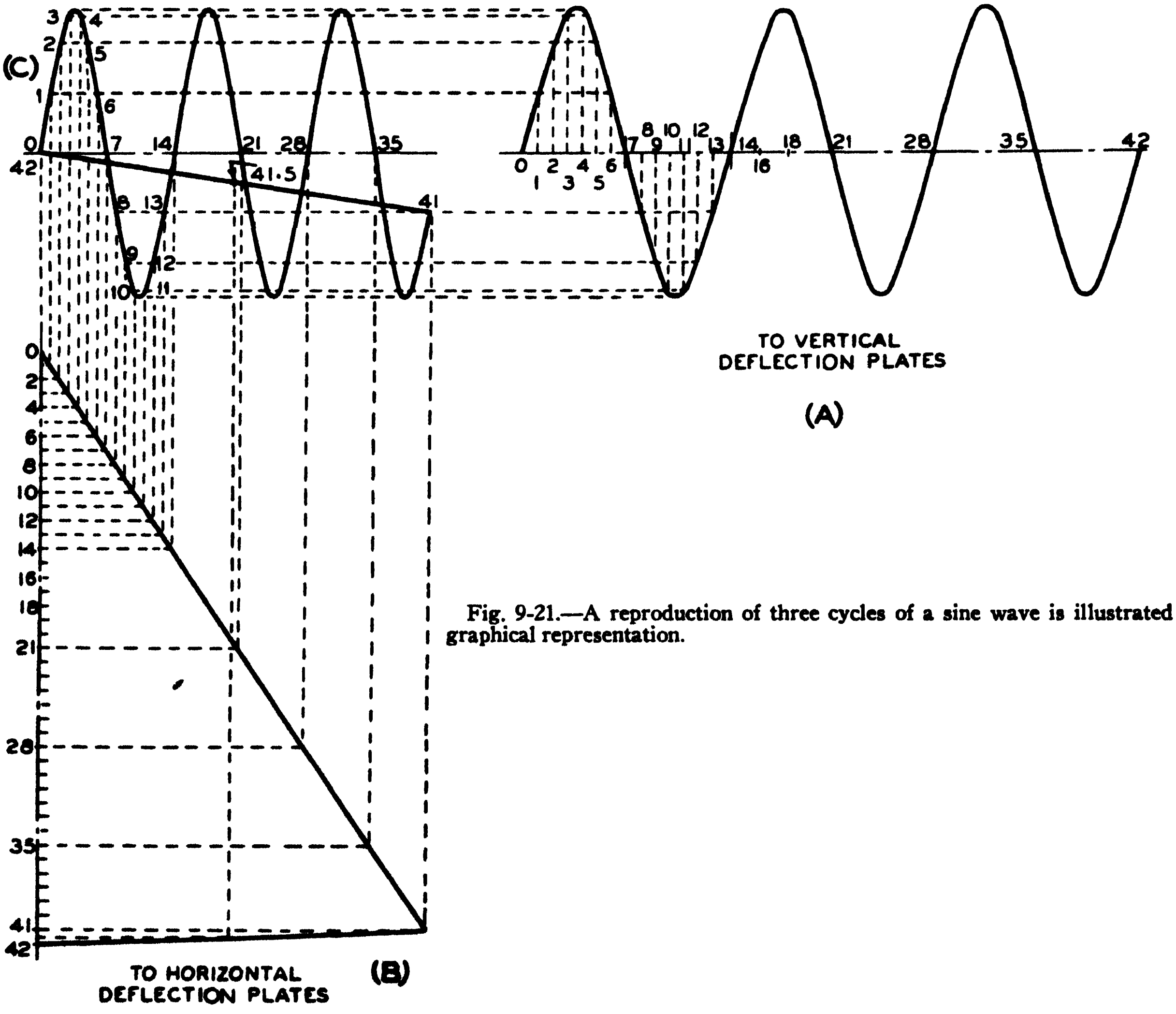

Fig. 9-21.—A reproduction of three cycles of a sine wave is illustrated in this graphical representation.

The interesting thing about these graphical illustrations is the effect of the retrace time of the sawtooth wave on the reproduced picture. In Figs. 9-19 and 9-20, the retrace time is 14.3 per cent of the total time of one cycle of sawtooth. This means that for each case, the reproduced sine wave will only be 85.7 per cent of the complete input wave. In Fig. 9-19, the flyback time of the sawtooth wave is from point *12* to point *14* along the time axis, which means that the part of the reproduced picture (C) between points *12* and *14* represents the effects of flyback time. In Fig. 9-20, points *24* to *28* indicate the flyback time of the sawtooth wave and the picture (C) from points *24* to *28* represents the effect of the flyback time. The two reproduced waves at (C) of both figures indicate that the flyback time makes the wave appear somewhat distorted, besides causing incomplete reproduction of the input sine wave.

In Fig. 9-21, the flyback time of the sawtooth is approximately 2.4 per cent of the total time of one cycle. This means that only 97.6 per cent of the input sine wave will be reproduced. The retrace time is from points *41* to *42* as seen in parts (A) and (C) of the drawing. From these three figures we can see that the shorter the retrace time of the sawtooth, the more complete and less distorted looking will be the reproduced patterns.

Let us now examine the drawings of Fig. 9-22, which are chosen to graphically illustrate what happens when the sawtooth sweep is nonlinear. The method of obtaining the reproduced curve is exactly the same as in the other figures. The number on the drawings corresponds to the same time lapse in each case. As mentioned previously, the effect of the nonlinear sweep is to cause uneven cycles to appear upon the screen, as seen in Fig. 9-22C.

Although three cycles of signal are shown in this case, you can easily imagine the effect of a nonlinear sweep upon a single- or two-cycle image. In a single-cycle image one of the half-cycles would appear normal

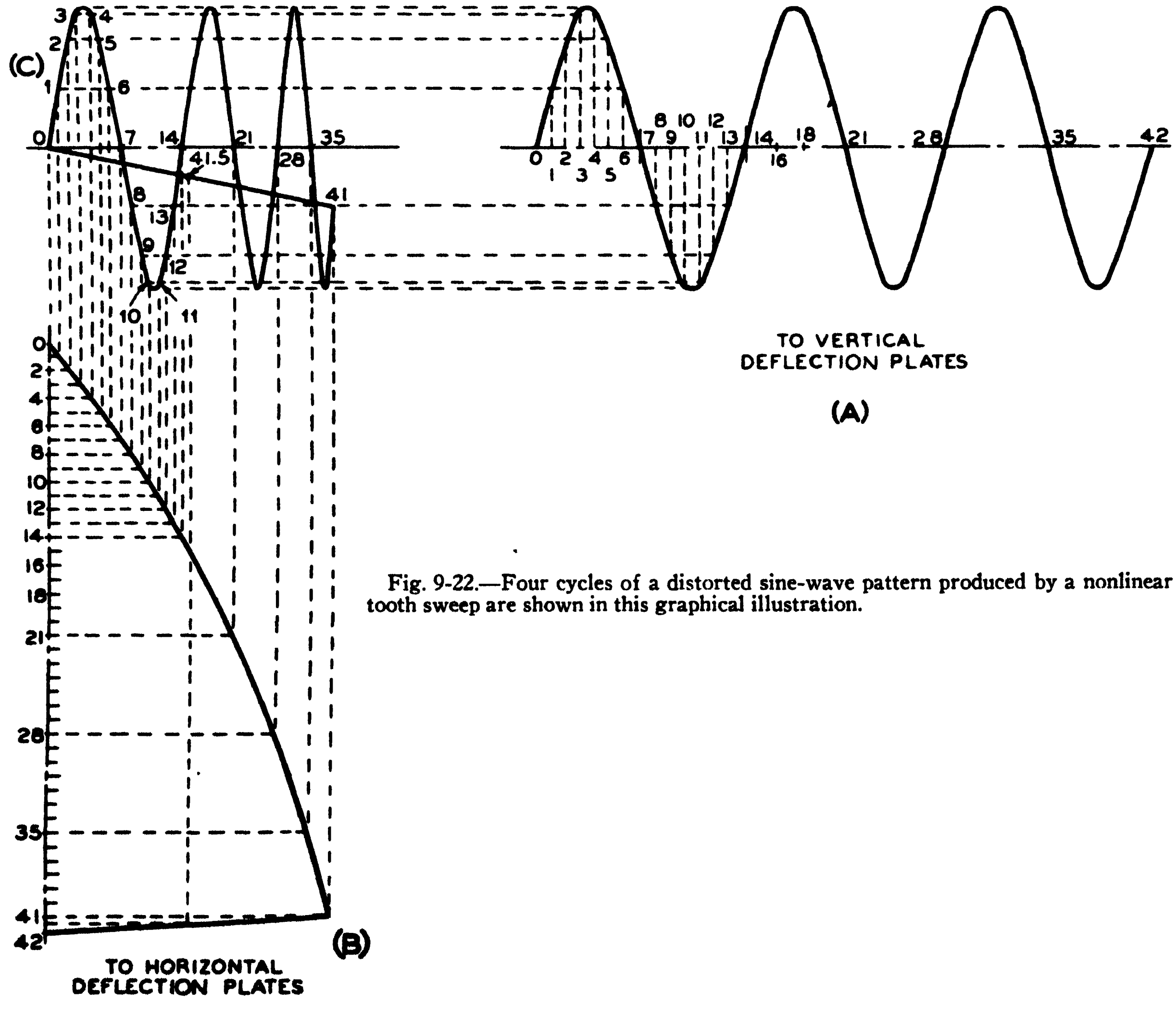

Fig. 9-22.—Four cycles of a distorted sine-wave pattern produced by a nonlinear sawtooth sweep are shown in this graphical illustration.

and the other half-cycle would be crowded. In the two-cycle pattern, one of the cycles would appear normal, whereas the other would be crowded. However, a nonlinear sweep may permit sufficient observation of the character of a waveform, if enough cycles are shown upon the screen. This is so, because, as you note, one portion of the nonlinear sweep voltage is a fair approximation of linearity. Hence with three or four cycles upon the screen, those close to the position which is the start of the beam may be considered as fairly true representations of the waveform voltage applied to the vertical-deflection plates. The sweep-voltage frequency in Figs. 9-19 through 9-22 has been kept constant and the voltage being observed has been changed in frequency. In practice, to obtain a greater number of cycles upon the screen, for any one frequency input to the vertical-deflection plates, the sweep frequency is changed.

Although we do not definitely tie in the sweep voltage shown in Figs. 9-19 through 9-22 with the basic electromechanical sweep of Fig. 9-2, it is to be understood that the operation of electronic sweeps in tracing the patterns is similar to that which has been described for the electromechanical sweep consisting of the battery voltage and the potentiometer.

Triangular Sweep

Some of the first sweep circuits employed a triangular-shaped wave to produce the time base. A basic drawing of a triangular wave is illustrated in Fig. 9-23. When this wave is applied to the horizontal-deflection plates as a sweep signal, the horizontal displacement of the spot traces itself across the screen during the time the wave goes from *A* to *B*. The so-called flyback or retrace time, *B* to *C*, of the triangular wave is exactly the same as the trace or rise time. This means that the spot sweeps back across the screen to its initial starting position at the same rate of speed as it advanced.

A triangular waveshape as the sweep signal is not desirable because it will, only under certain conditions, reproduce the input waveform upon the screen of the cathode-ray tube. The drawing of Fig. 9-24 makes this somewhat clearer. In each part of this drawing, *A* represents the input sinusoidal signal applied to the vertical-deflection plates, *B* the triangular sweep signal applied across the horizontal-deflection plates, and *C* the resultant pattern on the screen. In part (1), the frequencies of the vertical and horizontal signals are the same and the resultant wave is an oval or elliptical-shaped pattern. In part (2), the frequency of the sweep signal is decreased by half (or the input signal frequency doubled) and the resultant pattern is in the form of a horizontally positioned figure eight.

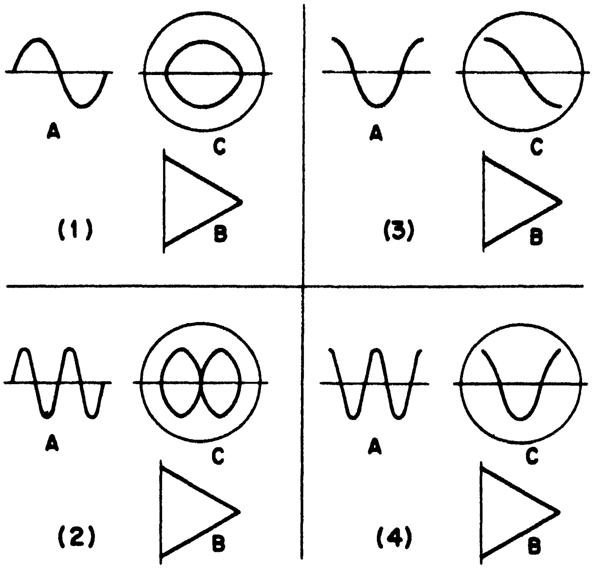

Fig. 9-24.—Drawings illustrating typical resultant patterns produced, if a triangular-shaped sweep signal is used. In each case, *A* shows the sinusoidal signal applied to the vertical-deflection plates, *B* the triangular sweep signal, and *C* the resultant pattern produced on the screen.

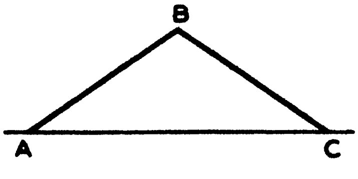

Fig. 9-23.—A triangular-shaped sweep signal used in some of the early time-base circuits.

In parts (3) and (4) of Fig. 9-24, the relative phase of the starting time of the two voltages has been changed. In part (3), where the frequencies of each are the same, the input signal has effectively advanced one-quarter of a cycle before the sweep takes effect. The resultant signal is seen to be a single line. In part (4), where the frequency of sweep is half that of the input sinusoidal signal, the sine wave is also advanced one-quarter of a cycle. The reproduced pattern in this latter case is seen to be an exact one-cycle reproduction of the input signal. Only one cycle appears because the same wave is double traced, therefore appearing as a single image.

It is the second half of the triangular sweep pattern that keeps this type of a waveform from being useful in sweep circuits. The sawtooth sweep is much better because its flyback time is negligible, as compared to the time required for the second half-cycle of the triangular sweep to be completed.

RELAXATION OSCILLATORS

In the following sections, we will study various types of electronic sweep generators. The major portion of the discussion will center around those circuits that are most commonly used to produce the sawtooth sweep voltage. Of these sweep circuits, the relaxation-oscillator type is one of the most commonly used in modern oscilloscopes, as well as in television receivers. The term relaxation oscillator has been used for quite some time and has taken on a number of meanings. In general, a relaxation oscillator is a generator of distorted waveshapes which has its oscillator action controlled by the charge and discharge of an inductor or capacitor through a resistor. A broad definition such as this encompasses many types of circuits.

There are three principle types of relaxation oscillators, namely gaseous-tube relaxation oscillators, multivibrators, and blocking oscillators. Gaseous tubes are often called soft tubes, while the multivibrator and blocking oscillator employ hard or vacuum-type tubes. There are other relaxation-oscillator circuits of the hard-tube type but the two mentioned above are the most common. In this section, we will study these three types of relaxation-oscillator circuits. In practice, when the term relaxation oscillator is used, it is often only associated with the gaseous-type or multivibrator circuit. The blocking oscillator, although fundamentally a relaxation oscillator, is not often thought of as such. Practically all of the relaxation oscillators that are used in oscilloscopes and in television receivers are of the capacitor-resistor variety; these are the types which will be discussed.

In the definition of relaxation oscillators, we stated that they are generators of distorted waves. This is in contrast with the normal signal generator, wherein anything but distorted waves are desired. Furthermore, the normal electron-tube oscillator develops a fundamental frequency determined by the values of capacitance, inductance, and resistance in the circuit and the order of the harmonics present is kept as low as possible. The relaxation oscillator's frequency, however, depends primarily upon the product of the capacitance and the resistance, with high harmonic content usually desired. Generally speaking, the frequency generated by such relaxation oscillators is increased as the capacitance and resistance over which the charge and discharge occur are decreased, and is decreased as the capacitance and resistance are increased. Similar variations in fundamental frequency and, naturally, the harmonic frequencies, can be developed, by varying but one of the constants, with the other maintained at a constant value. The frequency varies inversely with the capacitance and resistance. The exact frequency of oscillation depends upon other circuit factors also. This will be seen when the individual relaxation oscillators are discussed.

There are several reasons why such relaxation oscillators are better suited as time bases or sweep voltages than other types of oscillators. One of the chief reasons, as mentioned previously, is that they can produce a sawtooth wave which gives us a fairly linear time base with a rapid return trace. The brilliance of the return trace depends upon the fundamental frequency, for if this frequency is very high, the return of the spot, while much faster than the original sweep of the spot across the screen, may be sufficiently slow to produce a trace. A supplementary reason is that a fairly wide range of fundamental frequencies may be had with comparative ease and also that synchronization between two frequencies is easily accomplished. In other words, relaxation oscillators can be stabilized and kept constant by frequencies which are multiples or submultiples of the fundamental. This topic of synchronization is discussed in greater detail in Chapter 11.

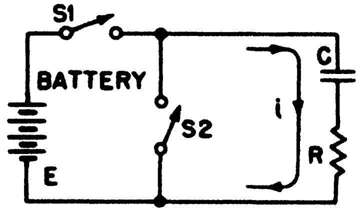

Fig. 9-25. — A series *R-C* circuit for charging and discharging capacitor *C* by means of switches *S1* and *S2*, respectively.

Capacitor Charge and Discharge Curves

Since most of the sawtooth relaxation oscillators employed in oscilloscopes utilize the basic charging and discharging effects of a capacitor through a resistor to produce their necessary waveshapes, we will study these developments first. In Fig. 9-25 is illustrated a series *R-C* circuit in conjunction with a battery and two switches. When switch *S1* is closed and *S2* is open, the charging current of the capacitor through the resistor is initially at a maximum, and starts to fall to zero immediately thereafter. This decrease in charging current is at an *exponential* rate, and if plotted, it will take the shape illustrated in Fig. 9-26.

From this transient curve we see that the maximum charging current, which occurs immediately at the time switch *S1* is closed, is equal to the value of the battery voltage *E* divided by the resistance *R*. After a period of time, the charging current decreases and approaches zero.[2] The charging current at the very start of its cycle of charge is considered to be quite linear, and if it decreased in this linear fashion, the curve would follow

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[2] The equation of the charging current is given by the expression

$$i = \frac{E}{R}\epsilon^{-t/RC}$$

where ϵ has a value approximately equal to 2.718. With known values of ϵ, *R*, and *C*, the curve of Fig. 9-26 can be plotted by substituting different values of time for t and solving for i.

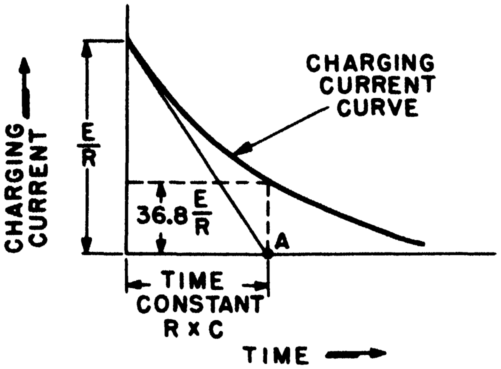

Fig.9-26.—Charging current curve of the capacitor showing the exponentially decreasing rate with time.

the path of the straight line as seen in the drawing of Fig. 9-26. When the *charging time* of the actual curve has reached point *A*, which is the zero-voltage point of the theoretical linear charge curve, the amplitude of the charging current will have decreased about 63.2 per cent of its maximum *E/R* value. This means that the charging current at this time is equal to 36.8 per cent of *E/R*. All these values can be demonstrated mathematically, but it is beyond the scope of this book.

At the same time the charging current is decreasing, a voltage charge *builds up* across the capacitor. This voltage charge increases until it reaches the voltage of the battery. The charge on a capacitor, after what is considered an infinite amount of time, is equal to the voltage applied, multiplied by the value of the capacitor. Expressed mathematically, we have $Q = CE$, where Q is the charge, C the capacitance, and E the voltage. Consequently, the voltage E across the capacitor is equal to Q/C.

We mentioned that this will occur after an infinite amount of time. Actually, as far as measurements with the most sensitive instruments are concerned, the potential value of Q/C is considered to occur within a finite value of time. The charging voltage curve is also an exponential function but a *rising* one. A graph of the charging voltage across the capacitor of the series *R-C* circuit of Fig. 9-25 is shown in Fig. 9-27. The charging voltage at the instant switch *S1* is closed is zero, and gradually rises to a value equal to *E*, the voltage of the battery.

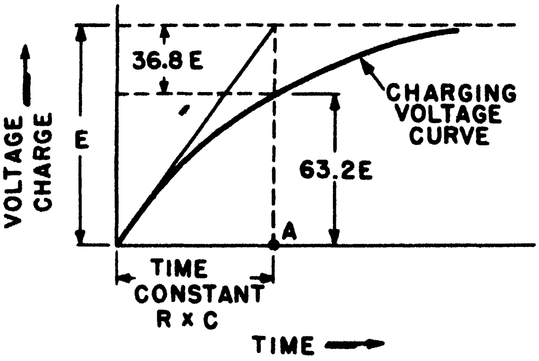

Fig. 9-27.—Graph of charging voltage across the capacitor.

If the charging curve continued to advance at its initial rate of rise, which is considered to be linear, it would assume the shape of the straight line in the drawing. In the time that it takes this theoretical linear curve to reach the maximum voltage of *E*, the actual voltage curve, which has an exponential rise, will have reached 63.2 per cent of *E*.

You will note that along the time axis of both the charging current and voltage curves, Figs. 9-26 and 9-27 respectively, the time to point *A* is marked off with the words "time constant." The ***time constant*** is the name usually applied to circuits to indicate certain times of charge and discharge. In the circuit under discussion, the time constant is an indicator of how long it takes for the charging current to decrease to a certain value or the charging voltage to increase to a certain value. The time constant for the circuit of Fig. 9-25 is equal to the time that it takes each straight-line theoretical curve of Figs. 9-26 and 9-27 to reach its final value.

The value of the time constant *t*, in seconds, can be computed from the product of the resistance of *R* in ohms by the capacitance of *C* in farads.[8] Thus

$$t = R \times C.$$

From these charging curves, we see that the time constant is an indication of how soon the charging current will decrease to 36.8 per cent of its maximum value or how soon the capacitor charging voltage will increase to 63.2 per cent of its maximum value. Certain multiple factors of the value of time constant also tell us a few things about the magnitude of charging current or voltage. In a time equal to twice the value of the time constant, the charging voltage will have increased or current decreased to 86.5 per cent of its maximum value; in a period of three times the time constant, the voltage will have increased or current decreased to about 95 per cent of its maximum value; and in four times the value of the time constant, over 98 per cent of the maximum will have been reached. For all practical purposes, we can consider that the maximum value of an exponential curve has been reached after five time constants have passed.

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[8] If the capacitor *C* of Fig. 9-25 were replaced by an inductor *L*, this circuit would also have a time constant, which would be equal to *L/R*, where *L* is in henrys, *R* in ohms, and the time constant in seconds.

Let us refer back to Fig. 9-25. Switch *S1* is closed, *S2* is open, and the capacitor is fully charged to the potential *E*. What happens now if switch *S1* is opened? First of all, the battery voltage is no longer effective in the circuit. Secondly, the charge on the capacitor is maintained, since there is no circuit through which it may leak off, However, if switch *S2* is closed, the capacitor discharges through resistance *R*. Curves representing the discharge of the capacitor are also of exponential form and have the same time constant. The interesting thing about this discharge sequence is that the current discharge curve is identical to the current charge curve. In other words, the decay of current is such that at the moment switch *S2* is closed, the current is maximum and decreases exponentially.

The discharge of voltage is such that the electric charge decays at the same rate as it increased, because of the same time constant. The voltage discharge curve appears exactly the same as the current curves; that is, they are decaying curves, which is opposite to the rising characteristic of the voltage charge curve. The factors of time constant also apply to the discharge curves as well as the charge curves. A drawing of the voltage discharge curve is illustrated in Fig. 9-28. Note that the time constant here indicates the time it will take the capacitor to discharge 63.2 per cent of its maximum voltage. To put it another way, the time constant indicates when the voltage will have decreased to 36.8 per cent of its maximum value.

The exponential curves of voltage charge and discharge of the capacitor are combined together in Fig. 9-29 to illustrate their over-all waveshape when the time constant of charge and discharge is the same.[4] In this drawing we assume that there is no initial charge on the capacitor during the period when switch *S1* of Fig. 9-25 is closed and that the capacitor has just become fully charged at the time switch *S1* is opened and *S2* closed. If switch *S2* were placed across the capacitor instead of across the battery, a completely different type of discharge curve would occur because the resistance *R* would no longer be in the discharge path. If the capacitor ideally had no leakage resistance, the discharge curve would be a vertical line similar to that in Fig. 9-16A. This presupposes a zero-value time constant for discharge, and is an important factor to remember, because in actual operation of relaxation oscillators, the tube effectively acts as a short across the capacitor during its discharge. Thus, in the succeeding sections on the discussion of sweep circuits, the capacitor alone, not the capacitor and charging resistance, will be effectively shorted.

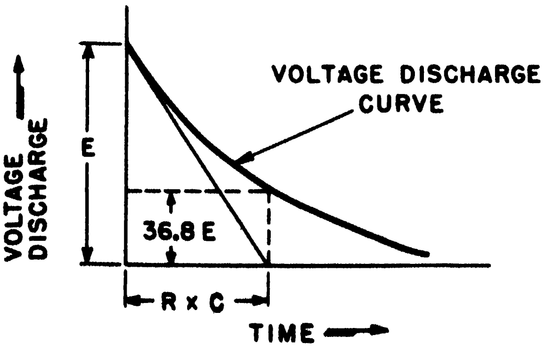

Fig. 9-28.—Voltage discharge curve of the capacitor.

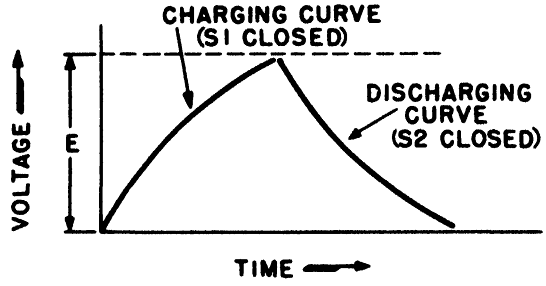

Fig. 9-29.—Combined charge and discharge curve of the capacitor showing over-all waveshape when charge and discharge time constants are the same.

A point of information relative to the shape of the exponential curves of charge and discharge can now be made. If the circuit time constant of Fig. 9-25 is increased or decreased, the slope of the exponential curves of charge and discharge will also change. This is so because the time constant always represents how soon the curve in question will increase or decrease to 63.2 per cent of its maximum amplitude. If the time constant were to remain the same but the supply voltage varied, then the slope of the exponential curve will also change because, in each case, the curve must reach 63.2 per cent of its maximum voltage in the same amount of time represented by the product of *R* and *C*. This will become more evident later in this chapter.

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[4] To be more exact, we should deal with charge and discharge curves of the electric charge *Q* instead of voltage. However, since the voltage is directly proportional to the electric charge, the curves apply equally as well to the voltage.

With this basic understanding of how a capacitor functions relative to its curves of voltage charge and discharge, we are now ready to study relaxation-oscillator circuits that make direct use of this principle in establishing a sawtooth sweep voltage.

Glow-Tube Relaxation Oscillator

In order to use the charging and discharging effects of a capacitor to help produce a sawtooth sweep, some means must be available whereby the capacitor can be *periodically* charged and discharged. The discharge, however, must occupy a very small part of the complete cycle of sawtooth. One of the first types of electronic circuits to employ this principle used a glow tube in a relaxation-oscillator circuit.

Glow tubes, often called glow-discharge tubes, are of special design in which the air has been evacuated and a small amount of gas inserted. Such tubes do not fall into the so-called vacuum-tube classification because of the gas they contain. For this reason, they are termed "soft tubes," distinguishing them from the "hard-tube" classification which is applied to vacuum tubes.

This type of tube requires a certain voltage to be applied across it to start it functioning. When this voltage is reached, the gas inside the tube *ionizes*. By this we mean that the molecules of gas inside the tube are converted into electrified particles, called *ions*. These ions make the tube conductive. When this period of ionization, also called the *breakdown* or *ignition* period, first occurs, it will cause the tube to flash and then start glowing. This glow will persist for a while even if the potential across the tube drops. However, when this potential drops below a certain value, the ionization ceases, and the tube again becomes nonconductive. Once this happens, the tube will again require the original voltage to cause ionization and hence conduction.

Thus we see that glow tubes have two values of critical voltage, namely that upper value causing ionization, called the *ionization potential, breakdown potential* or *firing voltage*, and the lower value that is the minimum required to maintain ionization of the gas and is sometimes referred to as *deionizing potential* or *extinction voltage*.

With this basic understanding of how a glow tube operates, let us now study an elementary relaxation-oscillator circuit, using a neon-type glow tube. The circuit for this is shown in Fig. 9-30. The resistor *R* and capacitor *C* are connected in series across the d-c voltage source and the neon tube is connected *directly across* the capacitor. Current flows through the resistor, and the capacitor begins to charge up. The voltage across *C* starts to increase in exponential fashion as shown by its charging curve in Fig. 9-31. Normally, were it not for the presence of the neon tube, the d-c supply voltage would charge the capacitor *C* to the full supply value, but with the tube connected across the capacitor, this does not occur.

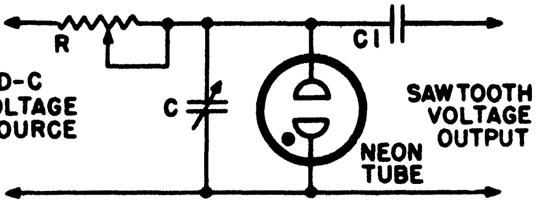

Fig. 9-30.—A typical relaxation oscillator using a neon-type glow tube.

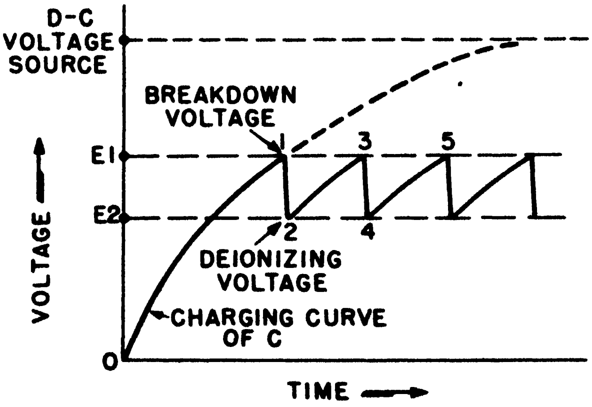

Fig. 9-31.—Voltage-time characteristic for the relaxation oscillator shown in Fig. 9-30.

What happens is that as soon as the charging voltage across the capacitor has reached the breakdown potential *E1* of the neon tube, the gas inside the tube ionizes and it begins to conduct. When this happens, and it is an instantaneous action, the tube effectively acts as a short circuit across the capacitor (its resistance during conduction is very low) and the capacitor immediately begins to discharge through the tube. If point *1* on the charging curve is assumed to occur at the breakdown potential of the neon tube, then at this point the capacitor *C* begins to discharge through the now-conducting neon tube.

As the capacitor discharges through the tube, the voltage across it will start decreasing. When this voltage decreases to the point where it is just equal to the deionizing potential *E2* of the neon tube, the tube stops conducting. The curve from points *1* to *2* in Fig. 9-31 indicates the discharge of the capacitor from the breakdown potential of the neon tube to the time when the tube becomes nonconducting. When point *2* has been reached, capacitor *C* of Fig. 9-30 stops discharging, because the low resistance path of the conducting neon tube no longer exists. A voltage charge still remains on *C,* however, equal to the breakdown potential *E2* of the neon tube. Since the d-c voltage source is still applied, the capacitor immediately starts to charge again. It starts charging, however, from point *2* because of the charge it still retains. The charging of the capacitor increases until voltage *E1* is reached once again, as shown at point *3* on the curve. At this voltage, the gas in the tube instantly becomes ionized and the tube once again conducts. Immediately, the capacitor stops charging and starts discharging through the neon tube once again. It discharges to point *4* on the curve, when *E2,* the lower potential at this point, causes the gas in the tube to deionize and, therefore, the tube becomes nonconducting.

From point *4,* the charge and discharge cycle of the capacitor continues as seen in the drawing. The curve between the voltage values of *E1* and *E2* represents the sawtooth sweep output from the circuit of Fig. 9-30. In this circuit, *C1* is the output capacitor used to couple the sawtooth voltage to any desired circuit, and usually is about 0.5 *μf*. The difference in potential between *E1* and *E2,* the difference between the ionizing and deionizing potentials of the glow tube, represents the peak-to-peak amplitude of the sawtooth signal.

When a system of this type is used to supply the sweep voltage or time base, the charging voltage, from points *2* to *3, 4* to *5,* and so on, is that part of the sawtooth signal which sweeps the beam spot across the screen; the discharge parts of the signal, points *1* to *2, 3* to *4,* and so on, are the retrace voltages which return the spot to its initial starting position. Consequently, the output voltage waveform of such a circuit is, in effect, a picture of just what is happening during the charge and discharge of the capacitor. The glow-discharge tube serves the purpose of discharging the capacitor at a predetermined instant. Just what this instant is in the units which are used in actual practice is a matter of design. The discharge at a certain instant, with respect to the voltage across the capacitor, bears a definite relation, as will be shown, to the frequency of the sawtooth signal produced and to the degree of linearity.

Fig. 9-32 is an oscillogram of two cycles of the output voltage waveform from a neon-tube oscillator circuit similar to that illustrated in Fig. 9-30. You can consider the rising exponential line as the charging cycle, since it indicates the rise in voltage across the capacitor. The discharge trace is the steep drop from maximum to zero, and is barely visible. This pattern is for a 400-cps wave, so that the duration of each cycle is 1/400th of a second. You can, therefore, see that in this case the duration of the discharge period is extremely small with respect to the charge period.

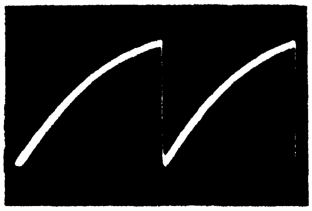

Fig. 9-32.—Oscillogram of two cycles of the output voltage from a neon-tube relaxation oscillator circuit.

If you examine Figs. 9-29, 9-31, and 9-32, you will note that the shapes of the discharge curves vary quite a bit. In Fig. 9-29, both the discharge curve and the charge curve are completed in the same amount of time, since the discharge and charge circuits have the same time constant. In Figs. 9-31 and 9-32, the period of discharge of the sawtooth wave is seen to be only a small part of the charging half-cycle. The discharging portions of the curves are not the same as the charging portions because the time constant of the circuit of Fig. 9-30 is different for the charge and discharge. This is a very important point to remember. Let us analyze this problem a little more thoroughly and see why this is so.

During the periods of charge, the neon tube is nonconducting and represents a very high resistance in the circuit as compared to the resistance of *R* in Fig. 9-30. Consequently, the time constant of the circuit during the charging of the capacitor is primarily determined by the product of *R* and *C,* the high resistance of the nonconducting neon tube having very little, if any, effect. The value of the time constant, as mentioned previously, determines the slope of the charging curve. When the ionization potential of the tube has been reached, the charging of the capacitor ceases at some point along its curve, the tube flashes and becomes conducting, and the capacitor starts to discharge through the tube.

At this point one may wonder why we say the capacitor discharges through the tube without making any reference to discharging current flowing through resistance *R.* The tube, which is directly across the capacitor, has the property of offering a very low resistance when the gas in the tube becomes ionized. Compared to the resistance of *R,* which in commercial sweep circuits usually has a minimum value of about 250,000 ohms, the resistance of the conducting tube is so low that it virtually appears as a short circuit across the capacitor. As a consequence of this short, the capacitor stops charging and starts discharging. The discharge current, naturally, takes the path of least resistance, which is through the tube.

The time constant of the circuit during discharge tells us how fast the capacitor will discharge and its effective resistance is primarily dependent upon the resistance of the tube during conduction plus the leakage resistance of capacitor C. Since the value of C remains the same during charge and discharge, the time constant during discharge is very much smaller than that during charge because in the latter case R is the effective resistance and is much greater than the total resistance during discharge. This smaller time constant means that the capacitor will discharge very rapidly compared to the rate of charge. Thus the resultant waveshape of the signal output as seen in Figs. 9-31 and 9-32 is in the form of a sawtooth signal with the discharging part of each cycle, which we term the flyback or retrace, but a small part of the complete cycle. As pointed out previously, this flyback time should be as short as possible because the greater this flyback time the less the amount of input signal reproduced. In other words, the discharge time constant of a relaxation-oscillator circuit should be as small as possible for complete reproduction of the input signal.

The frequency of a relaxation oscillator depends upon numerous factors. First of all, the values of resistor R and capacitor C are the primary determining factors. The d-c supply voltage, the ionization or breakdown potential, and the deionizing potentials are the other factors.[5] The frequency of the oscillator can, therefore, be varied by changing the value of any one of the above-mentioned constants. Let us consider variations in capacitance, resistance, and ionizing potential and see how they change the frequency of oscillations. With the breakdown voltage and the resistance constant, variation of the capacitor will vary the frequency. With the capacitance and breakdown potential constant, variation of the resistance will change the frequency. The frequency of oscillations decreases with increase of C or R. Actually, the frequency is inversely proportional to the product of C and R. This means that the higher the time constant of the circuit, the lower the frequency and the lower the time constant, the higher the frequency. The variation with respect to the breakdown voltage is such that the higher the breakdown or ionization or flashing voltage of the tube, the lower the frequency; and the lower the breakdown voltage, the higher the frequency.

It is relatively simple to understand why these variations in capacitance, resistance, and breakdown voltage vary the frequency. With fixed resistance, fixed supply voltage, and fixed capacitance, the time required to charge the capacitor to the breakdown voltage of the tube is finite. If the breakdown voltage is reduced, the time required to reach this value of charge or voltage across the capacitor will be less, hence the number of times the capacitor will reach that charge and discharge across the tube in a definite time will be greater, which means that the frequency will be higher. If the breakdown voltage is raised, then the period of time required to charge the capacitor to the critical voltage will be longer and the number of times the capacitor will reach that charge and be discharged by the tube in a definite period will be less and the frequency of the output voltage will be lower.

With a constant supply-voltage source, fixed resistor, and fixed breakdown voltage, the variation of frequency with capacitance variation is a function of time required to charge a capacitor to the proper voltage. The higher the value of capacitance, the greater is the time required to charge that capacitor to a certain voltage, and if this voltage is the critical ionization potential, the less will be the number of times that the capacitor will be properly charged and discharged in a definite amount of time. This means that the frequency will be lower. The smaller the capacitor, the less is the time required to charge it to a critical voltage and the more frequently will this capacitor reach its proper charge and be discharged in a definite amount of time. Hence, in this case, the frequency will be higher.

As far as the resistance is concerned, the greater the value of the resistor for fixed supply voltage, fixed capacitance, and fixed breakdown voltage, the longer is the time required to charge the capacitor to the critical voltage. Hence the number of charges and discharges is lower in a definite period of time; consequently, the frequency is lower. The lower the value of the resistance, the more rapidly will the capacitor receive its required charge in a finite time, and the greater the number of charges and discharges; hence the frequency is higher for a definite time interval.

What is said here is applicable to all such relaxation oscillators, even if they are more complicated than the

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[5] The frequency f of a relaxation oscillator similar to the circuit of Fig. 9-30, when based upon the assumption that the capacitor discharge time is negligible in comparison with its charge time, is given by the relation

$$f = \frac{1}{RC \log_e \dfrac{E - E2}{E - E1}}$$

where E is equal to the d-c supply voltage, $E1$ equal to the ionization or breakdown potential, and $E2$ equal to the deionization potential of the glow tube.

one illustrated in Fig. 9-30, and even if the fixed or variable resistance is replaced by some other device.

If you examine Fig. 9-32, you will notice certain significant details. Note that the portion of the cycle representing the rise in voltage or the charging period does not progress at a uniform rate with respect to time. That is, the rising voltage is not linear. The steepest portion is that which occurs as the charging cycle starts. The slope of the curve is greatest here. Then the slope becomes less and less until, when nearing the maximum amplitude (which means, approaching the critical voltage), the rise of the charge in the capacitor is very little for the few moments before the maximum is reached. If this complete rising voltage were applied to the horizontal-deflection plates so as to swing or sweep the spot across the screen, the speed of the spot would not be uniform with respect to intervals of time. The spot would move fastest at the start of its journey (which means at the start of the capacitor charging) and gradually slow down in its rate of travel as it approached the other limit of the screen. This was graphically illustrated in Fig. 9-18C.

A relaxation oscillator of this type can be used, but it would not be wholly satisfactory because it would develop a distorted image upon the screen. Nevertheless it has utility, because a certain portion of the charging cycle possesses sufficient linearity to permit examination of the pattern upon the screen. Provided that a number of cycles of the wave being observed are caused to appear upon the screen, the cycle or cycles positioned near the start of the spot travel would be substantially true images of the waveform being examined.

In connection with linearity of the sweep voltage, there are so many applications of the oscilloscope, where quantitative observations are not of prime importance, that a certain departure from perfect linearity is permissible. As a matter of fact, the commercial cathode-ray oscilloscopes now available possess sweep circuits which enable observations of various kinds and are of extreme value to the industry, yet which are not 100 per cent linear or not even approaching this value. Of course, all of these are more linear than the representation in Fig. 9-32, but they do have a certain departure from ideal linearity which is not sufficient to cause trouble.

The primary reason for the lack of linearity with the oscillator shown in Fig. 9-30 is that the use of a fixed or even variable resistor of the common type does not keep the flow of charging current at a constant rate; this requisite must be fulfilled if perfect linearity is to be attained. Let us consider a basic graphical analysis of how a sawtooth sweep can be made more linear. The analysis to follow is only theoretical. It assumes that the conditions of the relaxation oscillator are such that the supply voltage and time constant of the circuit remain the same and only the ionization potential of the tube in question can be varied.

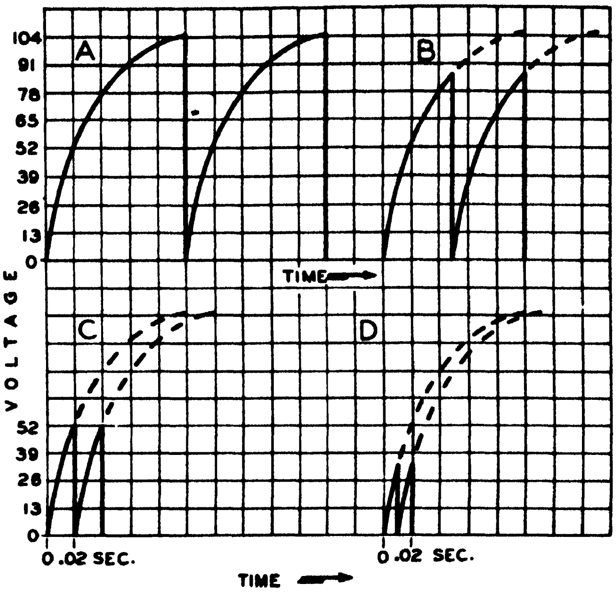


Fig. 9-33.—Appearance of sawtooth wave from a relaxation oscillator for different values of the ionizing potential.

The four parts of Fig. 9-33 illustrate the appearance of the sawtooth wave under different values of ionizing potentials. Curve *A* shows two cycles of a sawtooth wave supposedly developed by a fixed resistor-capacitor discharge tube combination, wherein the tube ionizes or flashes at 105 volts. The supply voltage is 110 volts. From what has been said, it is obvious that this type of sweep voltage would be unsatisfactory because the ionization potential of the tube is very close to the full supply voltage. Let us assume that the vertical lines indicate time along the horizontal direction where each division represents 0.02 second. A half-space, therefore, represents 0.01 second. For the first 0.01 second, the voltage rises approximately 33 volts. For the next it rises about an additional 20 volts. For the next, it rises by about 14.0 volts more. From the time of 0.03 second to 0.04 second the curve rises about 10.5 volts. For the next 0.01 second or 0.05 second after the start of the charge, it rises approximately 8 volts more, and at this point the charging voltage is equal to about 85.5 volts. For the remaining equal time variations of 0.01 second, this voltage rises in continually decreasing steps until the gas in the tube ionizes and the discharge occurs at 105 volts, which is 0.1 second after the start of the charge. (In this particular situation, we are assuming an ideal case, where the discharge is instantaneous from maximum to zero. In actual practice, a certain amount of time would elapse for the discharge period, because, as pointed out previously, the tube possesses resistance, even when it is ionized and the capacitor possesses leakage resistance. We are also assuming that the tube does not become a nonconductor, that is, has not become deionized, until the voltage has reached zero. In practice this is not so, as stated earlier in the reference to the upper and lower critical values of voltage.)

It is evident that the rate of charge and hence voltage increase is very irregular, actually being an exponential rise. Since the discharge occurs at 0.1 second after the start of the charge, and since we assume the discharge to be instantaneous, from maximum voltage to zero, then the frequency of curve *A* in Fig. 9-33 is 10 cps. Suppose we arrange the tube to ionize at 85.5 volts as shown at curve *B* of Fig. 9-33 and leave the charging supply voltage, resistance, and capacitance of the circuit the same. In such a case as this, the discharge will occur after an interval of 0.05 second. Hence the frequency is 20 cps. The effects of this reduction of ionizing potential, from 105 to 85.5 volts, besides increasing the frequency, will reduce the amplitude of the sawtooth signal and make the charge portion more linear than that at *A*. This means that the rate of travel of the spot across the screen will not vary as much as it did when the amplitude was greater as a result of using most of the charging voltage. It is to be understood that the rise in voltage with respect to time is not yet a close approach to being constant, but the change in speed from the start to the finish is not as great as before. From this we can gather a very important bit of information, namely, that if any attempt is made to employ the maximum amplitude by utilizing almost the full charging voltage, the departure from the required linearity is greatly increased. The change in frequency also brings to light the fact that under certain conditions, that is, if the adjustments of the sweep circuit change the amplitude very much, it is possible to cause a change in frequency of the sweep voltage.

Suppose that we reduce the breakdown voltage still more, say to 53 volts, as in curve C of Fig. 9-33. The charge and discharge period now occurs in 0.02 second. This means that the frequency of the sweep is 50 cps. It is obvious that the slope of the voltage curve is a closer approach to the ideal straight line than the curves of either *A* or *B*. The amplitude of wave C is further reduced.

Suppose that we reduce the ionization voltage still more, say to 33 volts, as shown in curve *D*. The charge and discharge cycle occurs in 0.01 second, hence the frequency has been increased to 100 cps. At the same time, the amplitude has been decreased still further. However, we are producing what appears to be a fairly linear wave. At no time is it possible to produce a truly linear wave, unless the control of the charging current is such that the flow is kept at a constant rate during the entire charging cycle. We have seen how, by utilizing a certain portion of the complete charging voltage to constitute the charging cycle, we can definitely improve the degree of linearity.

It might be well at this time if we made mention of the fact that the voltage values mentioned in connection with Fig. 9-33 were purely arbitrary. They do not designate the actual charging voltage values employed in commercial devices. The similarity, however, between these values and commercial equipment lies in the fact that the voltage used to charge the capacitor, through whatever current-limiting device is used, is much greater than the actual voltage developed across the capacitor and employed to ionize the gas-discharge tube.

Thyratron Triode Relaxation Oscillator

We have shown how the neon-type glow tube can be used as a sawtooth relaxation oscillator. Such a unit is, however, not suitable as the time base in oscilloscopes primarily because the ionization voltage of the tube cannot be controlled and, therefore, linearity and frequency could not be controlled. Many of the sawtooth sweep generators that are employed in oscilloscopes use a thyratron or gas-filled triode tube as the sweep oscillator. The basic principles of ionization and deionization are also effective here, but the tube grid allows some degree of control over the ionization potential of the tube.

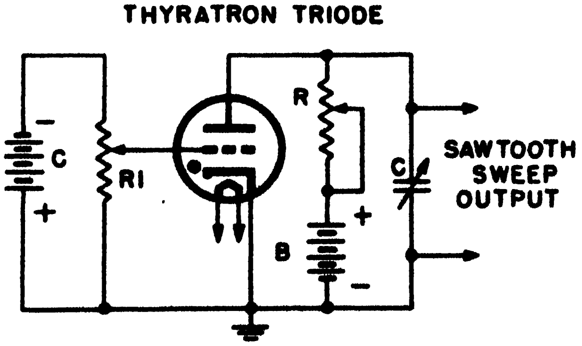


Fig. 9-34.—Basic circuit of a thyratron triode relaxation oscillator.

The thyratron triode consists of a plate, grid, cathode, and a separate heater element inside a glass envelope which is air evacuated but contains some inert gas. A typical triode thyratron is the type 884, which is used in many of today's oscilloscopes. A check of the sweep circuits employed by the oscilloscopes in Chapter 22 will indicate some of the types of gas triodes actually used by the different manufacturers.

A basic circuit of a gas-filled triode relaxation oscillator is illustrated in Fig. 9-34. The primary difference between this circuit and that of Fig. 9-30 is the addition of the grid circuit. The grid controls the ionization potential of the tube by controlling the number of electrons flowing from the cathode to the plate of the tube before the gas ionizes. With the bias on the grid at a certain potential, the voltage on the plate of the tube must reach a certain value before the gas in the tube will ionize. If the grid is made more negative, it will have a stronger control of the electrons emitted from the cathode and hence a higher plate potential is required to ionize the gas. On the other hand, the less negative the grid bias, the lower is the ionizing potential. Once the gas ionizes and the tube starts conducting, the grid loses control.

The basic cycle of operation of this relaxation oscillator circuit is very similar to that employing the neon tube. As battery *B* charges capacitor *C* through resistor *R,* the rising potential across this capacitor will, at some voltage determined by the grid bias, cause the gas in the tube to ionize and the tube will conduct. The tube then acts as a virtual short circuit across the capacitor. Thus, at the moment of ionization, the capacitor stops charging and starts discharging through the tube. The discharging will continue until the potential across the capacitor, and hence on the plate of the triode, reaches a certain value that will cause the gas in the tube to deionize. Once this point is reached, the tube stops conducting and appears as a high impedance across the capacitor; the capacitor once more begins to charge up from the *B* battery. As soon as the deionization occurs, the grid once more resumes control and its bias once again determines the ionization potential of the tube and, consequently, the start of the capacitor discharge.

In short, we find that the grid bias controls only the ionization or breakdown potential of the tube and not the deionization potential, and that the grid bias is only effective during the time when the gas in the tube is in the deionized state. In the drawing of Fig. 9-34, battery *C* is the bias supply and potentiometer *R1* controls the amount of bias applied to the grid of the tube. The sawtooth voltage output is taken across the capacitor. With this type of relaxation-oscillator circuit, the frequency of the sawtooth signal can now be varied by changing the values of any one of four factors, namely the bias on the grid, the capacitance of *C,* the resistance of *R,* and the voltage of the *B* battery. Since the bias on the tube controls only the ionizing potential, and since the deionizing potential remains the same, then variation in the bias, besides affecting the frequency of the sweep signal, will also vary the amplitude of the sweep signal. In practice, potentiometer *R1* is fixed, or least the bias voltage is fixed, and *R* and *C* are usually varied to produce the sweep voltage at the required frequencies. The value of fixed bias usually selected is the one that will produce the best linear sweep signal. Years ago, some commercial units did employ a potentiometer across a bias battery in order to provide an amplitude control of the sweep signal. Today, a major portion of the oscilloscopes use the power-supply voltage divider as the source of the bias voltage, and by properly tapping this source, the correct bias is provided.

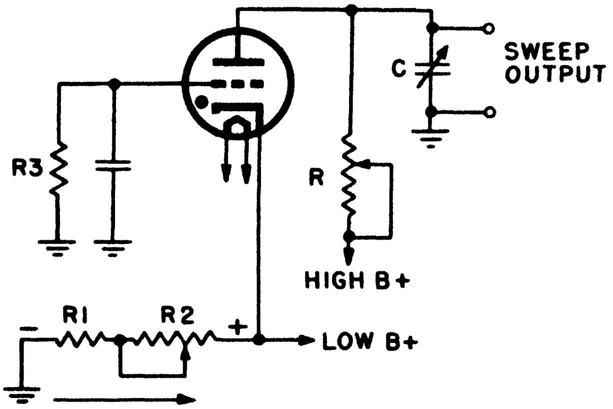


Fig. 9-35.—Simplified schematic of thyratron relaxation oscillator used in present-day oscilloscopes.

In some cases, a preset potentiometer is employed in the power-supply circuit so that adjustment of the bias voltage can always be made in case it is necessary. A simplified schematic of a relaxation-oscillator circuit as employed in oscilloscopes today is shown in Fig. 9-35. The cathode of the tube is connected directly to a low-voltage point in the power-supply circuit. The current flows in the supply circuit in the direction indicated and makes the cathode positive with respect to ground. The total resistance of *R1* plus the fixed value of potentiometer *R2* determines the bias on the tube. The total value of the current flowing in the supply circuit determines the choice of the total resistance of *R1* and *R2* because the voltage drop across these two resistors is the effective bias on the tube. Resistor *R1* is inserted to prevent the cathode from being grounded

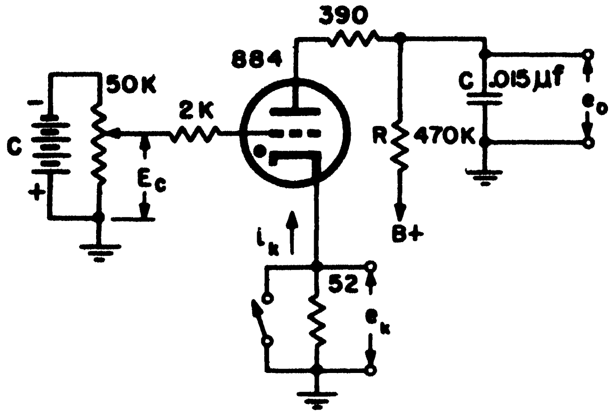


Fig. 9-36.—Experimental circuit of a relaxation oscillator using a type 884 gaseous discharge tube.

when the potentiometer *R2* is shorted out of the circuit. Thus, resistor *R1* assures that there will be some voltage on the cathode at all times. Resistor *R3* in the grid circuit completes the grid-return path to ground. The grid is at ground potential, and since the cathode is at a positive voltage, the grid is negative with respect to cathode.

An experimental circuit using a type 884 gaseous-discharge tube in the form of a relaxation oscillator is illustrated in Fig. 9-36. This circuit is typical of those employed in many oscilloscopes, except that the bias here is obtained from a battery. In this circuit, measurements were made on the sawtooth voltage output, the cathode current, and the frequency of the sawtooth for different values of grid bias and plate-supply voltage. For each measurement, pictures were taken of either the sawtooth waveform or cathode current, or both. In Table I, the different measurements are shown in tabulated form. Symbol E_b represents the plate-supply voltage, E_c is the grid bias, e_o is the peak-to-peak sawtooth voltage, e_k is the peak-to-peak voltage across the cathode resistor, i_k is the peak-to-peak cathode current, and f is the frequency of the sawtooth output.

TABLE I

Step	E_b volts	E_g volts	e_o volts	e_k volts	i_k ma	f cps
1	300	− 5	35	2	38	1,150
2	300	−10	90	4.7	90	360
3	300	−20	200	16	310	90
4	100	− 5	35	2	38	280
5	500	− 5	35	2	38	1,900

The 52-ohm resistor in the cathode circuit of Fig. 9-36 was inserted so that the cathode current could be calculated from the measured value of the voltage drop across this resistor. (It should be remembered that this calculated current is actually the discharge current of the 0.015-μf capacitor.) This resistor also enables us to observe the waveform of the discharge current by simply observing the voltage waveform across the resistor. In this case, the voltage waveform observed is the same as the current waveform because the voltage across and the current flowing through a resistor are in phase with each other. The cathode resistor had very little effect upon the operation of the circuit; nevertheless, it was shorted out when the different waveform patterns were taken. The switch across this resistor in the circuit of Fig. 9-36 is for this purpose.

The 390-ohm resistor in the plate circuit is usually found in most types of gaseous triode sawtooth relaxation oscillators. Its purpose is to limit the amount of discharge current that flows from capacitor *C* through the tube to the maximum allowable for the proper operation of the tube without harming it. The value of this resistor depends primarily upon the type of tube used and also upon the maximum amount of discharge current. Components *R* and *C* determine the time constant of the circuit.

For steps 1, 2, and 3 in Table I, the plate-supply voltage was held constant at 300 volts and the grid bias varied. It is seen that as the negative grid voltage is increased, the output sawtooth voltage is likewise increased. This establishes the fact that the variable potentiometer in the grid circuit of Fig. 9-36 acts as an amplitude control on the output signal. Besides the output signal, the cathode current is also increased as indicated by the increased voltage drop across the cathode resistor as the bias is made more negative. An increase in grid bias means that the ionization potential of the tube is raised. This, in turn, means that when the tube ionizes and causes the capacitor to discharge, it will do so when the charge on the capacitor has reached a higher value. Consequently, the discharging current will be higher, as indicated by i_k in Table I.

Measurement of the sweep frequency was made for each value of grid bias and was found to decrease as the grid bias was increased. The three photographs, (A), (B), and (C) of Fig. 9-37, each indicate the sawtooth output waveform taken across capacitor *C* for steps 1, 2, and 3, respectively, as the grid bias is increased from −5 volts to −10, and then to −20 volts. These patterns are not intended to indicate the frequency or amplitude difference of the sawtooth output but only to compare the linearity of the waves. The oscilloscope upon which these waveforms were observed was adjusted to produce a pattern of three cycles for each picture with about the same amplitude so that the best comparative analysis of the linearity

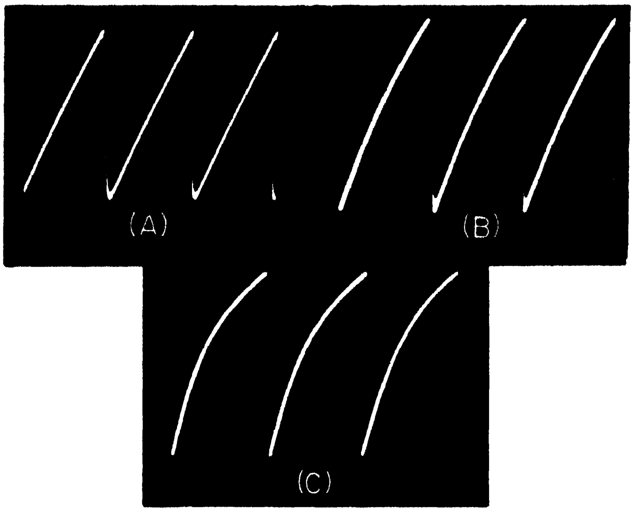


Fig. 9-37.—The sawtooth output waveform is shown for the circuit illustrated in Fig. 9-36. The grid bias is progressively increased from −5 volts in (A), to −10 volts in (B), and to −20 volts in (C).

could be made. From these patterns, it is readily concluded that as the grid bias is increased beyond what is termed its safe operating value, the sawtooth wave becomes nonlinear. The more negative the bias, the greater the nonlinearity in the sawtooth.

For step 4, the grid bias was returned to its normal voltage of −5 volts and the plate supply reduced to 100 volts. By comparing steps 1 and 4, you will note that changing the plate-supply voltage does not produce any change in the cathode current i_k, or sawtooth voltage output e_o, but only produces a change in the frequency of the sawtooth wave. This change in frequency is such that as the voltage is decreased, the frequency is also decreased; the decrease in the case under discussion is from 1,150 cps to 280 cps. The output sawtooth wave for step 4 is shown in Fig. 9-38. Upon comparison of this oscillogram and that of Fig. 9-37A, it is readily apparent that the former picture is nonlinear in contrast to the latter.

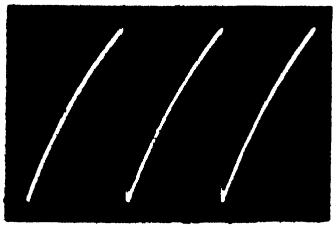


Fig. 9-38.— Output sawtooth wave is shown for the circuit of Fig. 9-37 after the grid bias has returned to its normal value of −5 volts and the plate supply voltage reduced to 100 volts.

In order to understand how a decrease in supply voltage will affect the frequency and linearity of the sawtooth waveform, let us study the charging curves of Fig. 9-39. Awhile back we mentioned that if the time constant of an *R-C* circuit remained the same and only the supply voltage changed, then the slope of the charging curve would also change. This is so because the time constant being the same for each curve means that the capacitor will become fully charged in the same amount of time, no matter what the supply voltage is. In Fig. 9-39, curve *A* represents the exponential charge on the capacitor for supply voltage *E1,* and curve *B* represents the charging curve for the lower supply voltage, *E2.* Note that each curve reaches its maximum voltage in the same amount of time. Since the grid bias is the same for each curve, then the gas tube of Fig. 9-36 will ionize at the same potential as long as the supply voltage is always higher than this ionizing potential. Therefore, in Fig. 9-39, *E3* represents the breakdown voltage of the tube and is seen to be the same for both curves. *E4* is the deionization potential of the tube and is likewise seen to be the same for each curve. Since the difference in voltage between the ionizing and deionizing potentials of a thyratron relaxation sawtooth oscillator determines the voltage of the sawtooth wave, then it is readily conceivable why the sawtooth output voltages for steps 1 and 4 of Table I are the same.

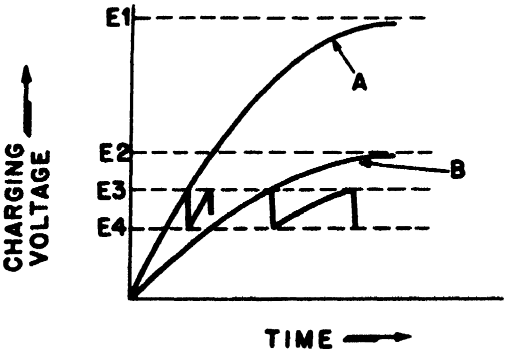


Fig. 9-39.—Voltage charging curves for different values of the supply voltage.

Let us now examine both sawtooth waves of Fig. 9-39 and see how this graphical analysis compares with the actual measurements and photographs taken. From the two curves, it is seen that curve *A* has a steeper slope than *B*. Since ionization for both curves occurs at the same time, the charging characteristic of the sawtooth wave produced by curve *A* will be more linear than that of *B*. This proves the point that if the supply voltage of a relaxation oscillator were lowered below normal, the new sawtooth wave would be nonlinear. Comparing curves *A* and *B* once more, it is noticed that for curve *A* the charge and discharge during the breakdown and deionizing potential of the tube occur at a faster rate than those for curve *B*. Thus we

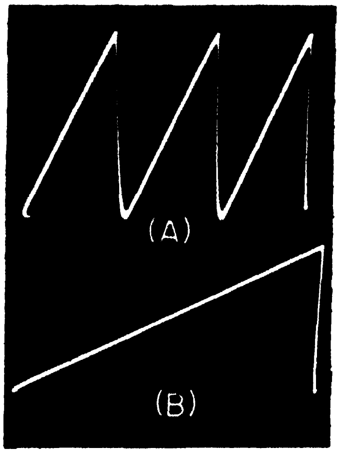


Fig. 9-40.—Sawtooth output waveform after the supply voltage has increased to 500 volts with the grid voltage remaining at −5 volts. (A) shows three cycles of the wave, while (B) illustrates one cycle occupying the same time-base length as the three cycles.

see that the frequency of the sawtooth produced by the wave at *B*, which is for the lower supply voltage, is lower than that of wave *A*.

In step 5 of Table I, the grid voltage remains at −5 volts, but the supply voltage is increased to 500 volts. Once again we see that the sawtooth output voltage, e_o, and the cathode current, i_k, remain the same as in step 1. The only thing that has changed is the frequency of the sawtooth signal, increasing from 1,150 cps, for a supply voltage of 300 volts, to 1,900 cps for the supply voltage of 500 volts. A picture of the sawtooth output waveform for step 5 is illustrated in Fig. 9-40. That photo in part (A) shows three cycles of the wave. Although not easily detectable, the sawtooth of Fig. 9-40A is more linear than that of Fig. 9-37A. The reason for this slight difference in linearity, even though the supply voltage has been increased by 200 volts from normal, is because there is not much difference in the slope of the charge curves between voltages *E3* and *E4* when the supply voltage is increased beyond what is considered its normal value of operation. In order to better illustrate the linearity of the sawtooth wave at (A) of Fig. 9-40, one cycle of this wave was made to occupy the same space as the three cycles and then photographed, as seen at (B) of the same figure.

An important characteristic of the circuit of Fig. 9-36 is the shape of the cathode current that flows during the discharge periods of the capacitor. Fig. 9-41 is an oscillogram of the cathode current that occurs for step 1 of Table I. The peak current of this curve is equal to 38 ma. Its frequency, 1,150 cps, is the same as the sawtooth for the conditions of step 1. The small gap in the base of the curve during each current pulse indicates the time of the current discharge of the capacitor, which occurs only between the ionization and deionization potentials of the tube. Since the flyback or retrace time of the sawtooth wave also occurs only during the discharge of the capacitor, then the flyback time of Fig. 9-37A and the period of each current pulse of Fig. 9-41 are the same. During the time between current pulses, the capacitor charges. By a measurement of the gap in the base of the curve of Fig. 9-41 and of the time interval between current pulses, the time ratio of discharge to charge of the capacitor can be determined.

In order to determine the effect of a higher value of the plate limiting resistor, the 390-ohm resistor of Fig. 9-36 was replaced by one of 10,000 ohms. The cathode resistor was changed to about 400 ohms in order to obtain a sufficient voltage drop across it for a picture of the cathode current. In order to sustain oscillations, the resistor *R* had to be increased to about 1.5 megohms. The *B*+ voltage was set at 300 volts and the grid bias at −5 volts. The measured results were as follows: $e_o = 18$ volts, $e_k = 1.2$ volts, $i_k = 3.1$ ma; the frequency of oscillation *f* was 600 cps.

The pictures of the sawtooth voltage output and the cathode current are indicated in parts (A) and (B), respectively, of Fig. 9-42. The sawtooth voltage at part (A) is not a very good sweep wave because the retrace time forms an appreciable part of the complete cycle. From the cathode current of part (B), we can readily measure the ratio of discharge to charge, which is about 5 to 18. Thus the discharge portion occupies

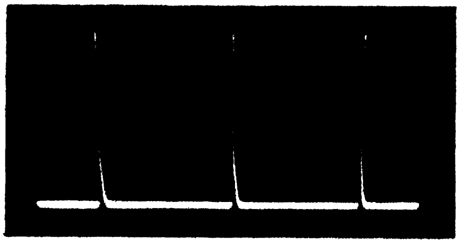


Fig. 9-41.—Oscillogram of the cathode current of the circuit of Fig. 9-36 under the conditions of step 1 in Table I.

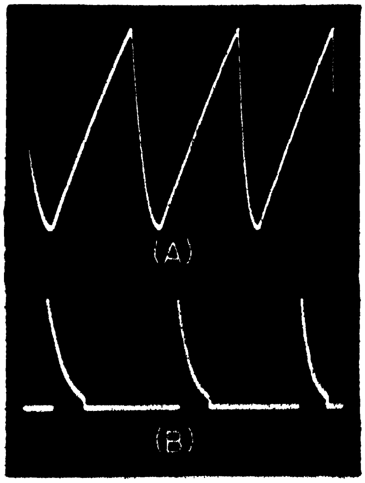


Fig. 9-42.—Oscillogram of the sawtooth voltage output is shown in (A) and that of the cathode current in (B).

nearly 22 per cent of the complete cycle which is definitely undesirable.

A point of interest relative to the oscillograms in Figs. 9-37 through 9-42 is that they do not show any return trace of the spot. What actually happened is that during the retrace time of the sweep that is employed in the oscilloscope from which the pictures were taken, the control grid of the cathode-ray tube was biased beyond cutoff. This biasing action prevents electrons from the cathode of the tube from reaching the screen. This effect is commonly referred to as "blanking" and will be discussed in detail in Chapter 10.

In the previous analysis, we indicated what happens to the sawtooth output of Fig. 9-36 with different values of supply voltage and grid bias. The normal operation is that represented by step 1 of Table I. In order to test the sweep operation of this circuit, the values of *R* and *C* were made variable. A 5,000-cps sine wave was applied to the vertical-deflection plates of an oscilloscope and the sawtooth output from the relaxation oscillator was applied to the horizontal plates to function as the sweep signal. With the normal operating potentials on the gas triode, the values of *R* and *C* were so adjusted that two cycles of the sine wave appeared on the screen of the oscilloscope.

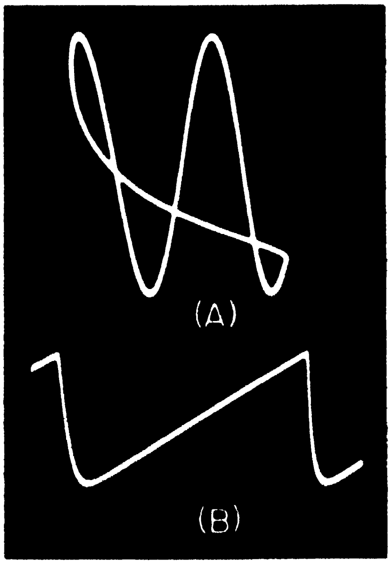


Fig. 9-43.—(A) shows the reproduction of two cycles of 5,000-cps sine wave applied to the vertical plates of the circuit of Fig. 9-36. The corresponding 2,500-cps sawtooth sweep signal is shown in (B).

The photograph of this reproduced sine wave is illustrated in Fig. 9-43A. The oscilloscope upon which this picture was taken did not employ blanking and thus the retrace of the spot is evident. Note that the two cycles are not complete and that the return trace occupies a fair part of the picture. In order to produce two cycles of a 5,000-cps sine wave, the frequency of the sweep, which in our case is that of the thyratron relaxation oscillator, should equal 2,500 cps, which is half that of the input signal. The sawtooth signal of 2,500 cps that reproduced the pattern of Fig. 9-43A appears in (B) of the same figure. Note that the flyback time of each cycle is not as short as might be desired. This is the reason why the sine wave at (A) is not complete reproduced. The photos of Fig. 9-43 should be compared with the line drawings at (B) and (C) of Fig. 9-20.

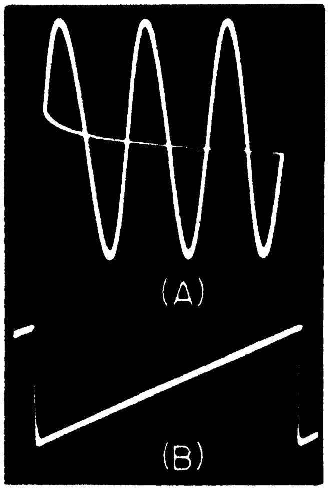


Fig. 9-44.—In (A) the *R* and *C* values of the sweep oscillator were adjusted to reproduce three cycles of a 150-cps sine wave applied to the vertical plates, while (B) shows the corresponding sawtooth sweep signal.

In another test, a 150-cps sine wave was applied to the vertical-deflection plates of an oscilloscope and the sawtooth output voltage from the oscillator of Fig. 9-36 applied across the horizontal-deflection plates. The *R* and *C* values of the sweep oscillator were so adjusted that three cycles of sine wave appeared on the oscilloscope screen. A photograph of this reproduced wave appears in part (A) of Fig. 9-44. This picture shows that the three cycles of sine wave are almost complete and that the return trace occupies only a small portion of the entire picture. This tells us that the sawtooth sweep signal must have a fairly short flyback time. A photograph of this sawtooth signal is shown in part (B) of Fig. 9-44. Note that the discharging part of each cycle has a sharp sloping characteristic. The frequency of the sawtooth wave is 50 cps because it reproduces only three cycles of the 150-cps sine-wave input. If the two photographs of Fig. 9-44 are compared with parts (A) and (B) of Fig. 9-21, they will be found to be quite similar.

If you will examine the photos of Figs. 9-43 and 9-44 once again, you will note that in each case the time occupied by the complete picture of the sine waves of part (A) is the same as that occupied by one cycle of the sawtooth sweep at (B). From the ratio of the number of cycles of the picture at (A) to its corresponding picture at (B) for the same amount of time, the frequency of either signal can be computed once the other one is known.

Checking Linearity of Sawtooth Sweep

At times it may be desirable to check the linearity of the sawtooth sweep signal that is used in many oscilloscopes. Of course, an exact picture of the sweep signal would be a fair indication of what the linearity would appear like, but it would be difficult to determine the degree of nonlinearity. Furthermore if the sweep of one oscilloscope is reproduced on another oscilloscope, we do not know if the latter unit itself has a linear sweep.

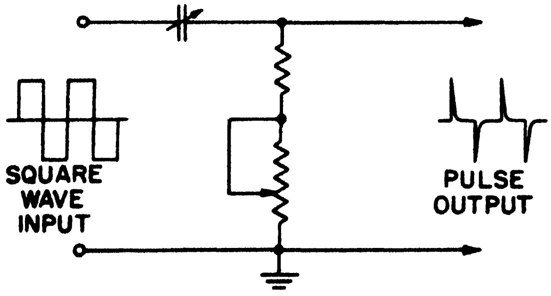


Fig. 9-45.—An *R-C* differentiating network showing the shape of the input and output pulses.

A quick and accurate method for checking sawtooth sweep linearity is to feed into the vertical plates of the cathode-ray tube a very sharp periodic pulse whose frequency is known. If the spaces between the pulses on the screen of the cathode-ray tube are equal, then the sweep is linear; if the pulse spacing is unequal, then the sweep is nonlinear. One method of obtaining such a pulse is to use a square wave and differentiate it. The time constant of the differentiating network determines the period of each pulse. The shorter the time constant, the sharper is the pulse. Fig. 9-45 illustrates the square-wave input to a differentiating network and the desired pulse output when the time constant is low. These pulses are then applied to the vertical terminals of the oscilloscope.

Without any input to the vertical plates, but with the sawtooth sweep signal applied to the horizontal plates, all that would appear on the screen of the oscilloscope would be a horizontal straight line as shown in Fig. 9-1. From this time base, we have no way of knowing how linear the sweep is. The pulses of Fig. 9-45 are equally distributed along their time axis. Consequently, if the sawtooth sweep is linear, when these pulses are applied across the vertical-deflection plates they will be reproduced in the same equally distributed fashion.

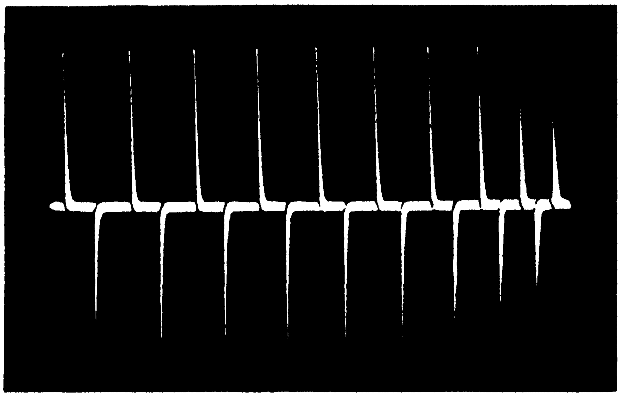


Fig. 9-46.—The unequal spacing of the pulse markers shown in this photograph indicates nonlinearity of the sweep at 50 cps.

In a simple experiment, we used a 500-cps square wave and produced a series of periodic pulses by the arrangement shown in Fig. 9-45. The pulse frequency is, therefore, also 500 cps. This pulse output was then applied to the vertical terminals of a laboratory oscilliscope that was in good working condition. The sweep frequency of the oscilloscope was set to about 50 cps in order to test the linearity of the sawtooth sweep of the oscilloscope at that frequency. With the 500-cps pulse input signal being 10 times greater in frequency than the sweep signal, approximately 10 cycles of the pulse appear on the screen. A photograph of how these pulse markers appear is illustrated in Fig. 9-46. The degree of nonlinearity of the sweep can be seen readily by examining either the positive or negative pulses. We do know that the pulse input is constant in frequency so the unequal spacing of the pulses in the photo indicates that the sweep of the oscilloscope is nonlinear at 50 cps. Note that the spreading of the pulses is greatest at the left side of the photo and bunched on the right side.

The ratio of the spacings between the first two positive peaks on the left to between the last two positive peaks on the right is about 5 to 2. This means that the charging cycle of the sawtooth sweep signal at 50 cps increases in amplitude at the very beginning of its charge at a rate two and a half times as fast as it does at the end of its charge.

Now let us examine the linearity of the sawtooth sweep from the same oscilloscope at a higher frequency. The frequency of the square-wave input to the differentiating network and hence the pulse frequency in this case was adjusted to 6,000 cps and the sawtooth sweep of the oscilloscope was set to 600 cps. The resultant pattern appears in Fig. 9-47. Examination of the pulse

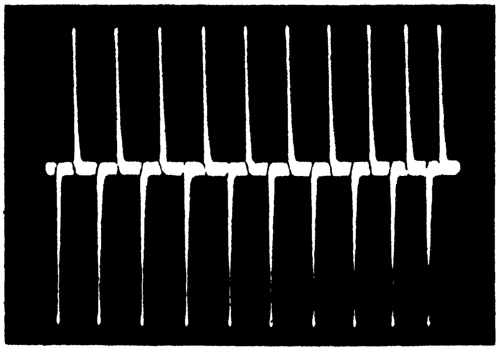


Fig. 9-47.—Here the pulse markers show relative linearity of the sweep frequency at 600 cps; compare this with Fig. 9-46.

markers on this photograph shows that the markers are more evenly spaced indicating that at 600 cps, the sweep is more linear than at 50 cps. There is, however, a very small degree of nonlinearity at the sweep frequency of 600 cps, but it will hardly be noticed in the reproduction of waveforms.

If the square wave or pulse of Fig. 9-45 were rectified before being applied to the oscilloscope, either the positive or negative half-cycles of the pulse would appear as markers. This would make checking the nonlinearity of the sweep easier to see. In the experiment conducted here, a diode rectifier was used, with the pulses applied to the plate of the diode. As a result of the circuit hookup, the negative pulse peaks were clipped and only the positive peaks remained. The experiments relative to Figs. 9-46 and 9-47 were repeated with the negative peaks of the pulse input clipped. The resulting positive markers are illustrated in Fig. 9-48A and B, respectively. Note how easy it is to check on the linearity of the sweep signal.

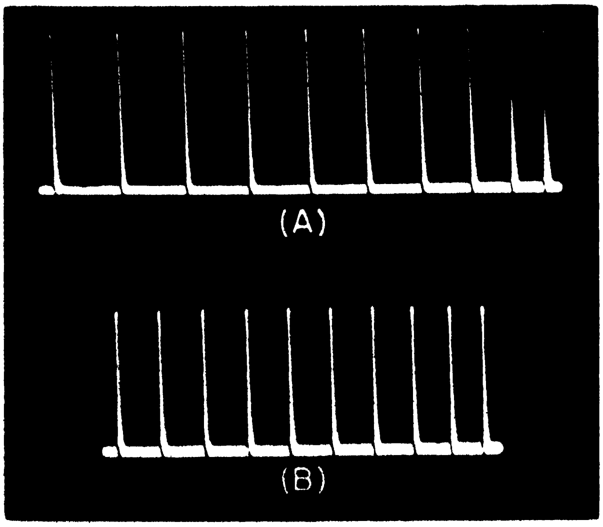


Fig. 9-48.—(A) and (B) show the pulse markers for the same conditions as in Figs. 9-46 and 9-47, respectively, but here the negative peaks of the pulse input were clipped.

Some of the commercial oscilloscopes have internal provision for a marker system similar to the method just discussed. For example, the Du Mont Type 248 (see Chapter 22) employs a special sine-wave oscillator which has its output signals clipped and amplified to the point where it will be a square wave. The square wave is then differentiated into sharp pulses. In this system the differentiated signal is clipped and then applied as strong negative pulses to the *grid* of the cathode-ray tube. The pulses bias the tube beyond cutoff and cause the beam to be blanked out periodically; hence blank spots appear along the pattern. The linearity is checked by determining the horizontal separation between blank spots. This oscilloscope, as well as most others that incorporate marker systems, has provision for changing the repetition rate of the marker pulse.

Practical Thyratron Sweep Circuits

Before we study the hard-tube circuits, let us examine a typical thyratron-triode sawtooth sweep circuit in its entirety as used in oscilloscopes today. The circuit is shown in Fig. 9-49. It will be noticed that this drawing is not much different from that of Fig. 9-35.

The bias necessary for the desired ionization potential of the tube is obtained from the B supply. The two cathode resistors, *R1* and *R2,* establish a positive charge on the cathode of the tube from the power supply with respect to ground. The grid resistor *R3* puts the grid at zero or ground potential. The grid is therefore, negative with respect to the cathode. Resistor *R4* is the discharge current-limiting resistor.

The frequency-determining elements are resistor *R* and *R5* and whatever capacitor is chosen by the switch. The switch represents the so-called *coarse frequency*

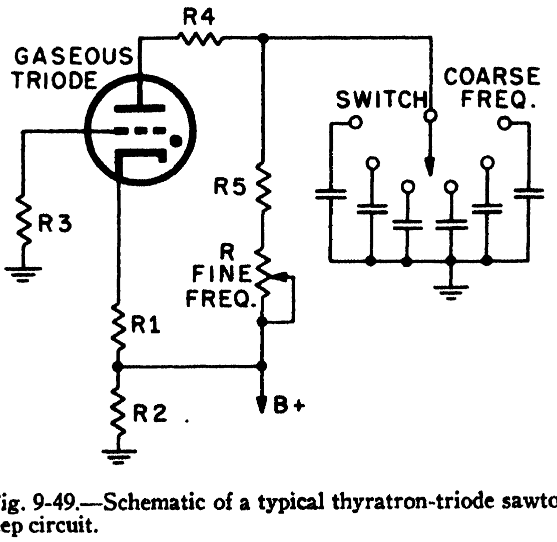


Fig. 9-49.—Schematic of a typical thyratron-triode sawtooth sweep circuit.

control of the sweep circuit. The potentiometer *R* operates in conjunction with the capacitor chosen, and selects the exact frequency of the sawtooth sweep. This latter adjustment is, therefore, termed the *fine frequency control.* The smaller the capacitor selected, the smaller is the time constant and the higher is the sweep frequency.

In this drawing, we have omitted the synchronizing circuit; nevertheless, one does exist in each oscilloscope. The topic of synchronization is discussed in detail in Chapter 11.

Multivibrators

The other form of relaxation oscillator that is employed in commercial cathode-ray oscilloscopes is the well-known multivibrator circuit. Multivibrator circuits are used a great deal in radar and television to produce different types of waveforms, one of the most common being the square wave. Such circuits can also be made to produce a sawtooth signal output that can be used as the linear sweep in oscilloscopes and television receivers. A number of multivibrator sweep circuits will be found in the schematic diagrams in the commercial cathode-ray-oscilloscope chapter.

A multivibrator employs two tubes, which may or may not be in a common envelope, and which are normally resistance-capacitance coupled to each other. There are many different types of multivibrators. It is beyond the scope of this book to completely discuss all the different types. Whatever is studied here will have some bearing upon the sweep circuits as used in cathode-ray oscilloscopes. A multivibrator can be made to operate by some outside driving source, or it can be run freely at a frequency dependent upon its own circuit arrangement. In general, multivibrators have two basic classifications, namely the *continuous, free running* or *recurrent* type, and the *triggered* or *driven* type; each of these classifications is subdivided into further types as will be seen later. Both types find application as sawtooth sweep generators in cathode-ray oscilloscopes as well as numerous other uses.

Multivibrator circuits, being a form of relaxation oscillator, make use of the charge and discharge of a capacitor through a resistance in producing its frequency of oscillation. The capacitor charge and discharge curves, as previously discussed, apply as well in the following analysis of different multivibrator circuits.

Basic Plate-Coupled Multivibrator

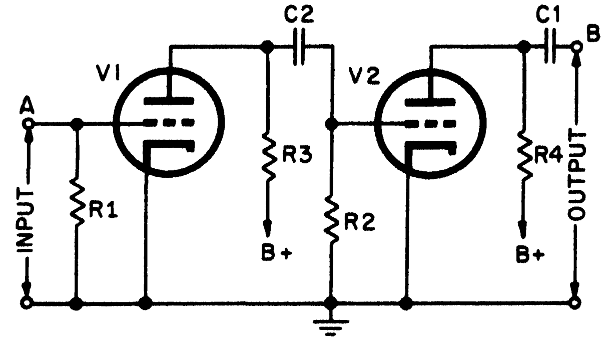


Fig. 9-50.—Circuit of a conventional *R-C* coupled two-stage amplifier.

The so-called original multivibrator circuit is essentially nothing more than a two-stage resistance-capacitance coupled amplifier employing feedback, where the output from the second stage is coupled back to the input of the first stage. In Fig. 9-50 we see a typical two-stage resistance-capacitance coupled amplifier indicating the input and output circuits. The components are standard, with *R1* and *R2* being the grid resistors of tubes *V1* and *V2* respectively, *R3* and *R4* being the plate-load resistors, and *C1* and *C2* being the coupling capacitors. If we take the output circuit at point *B* and connect it back to the input circuit at point *A,* we essentially have the basic plate-coupled multivibrator circuit. This type of circuit is illustrated in Fig. 9-51, and is drawn in the conventional manner; it is the same circuit as that of Fig. 9-50, with the connection from *B* to *A* completed. This circuit is of the free-running type.

Each stage of a resistance-capacitance coupled amplifier causes a 180° phase reversal in signal between its plate and grid circuit, and there is normally negligible phase shift by the *R-C* components between two stages. Thus the signal fed back from the plate circuit of *V2* in Fig. 9-51 to the grid circuit of *V1* is in phase with the signal already existing at the grid of *V1*. Con-

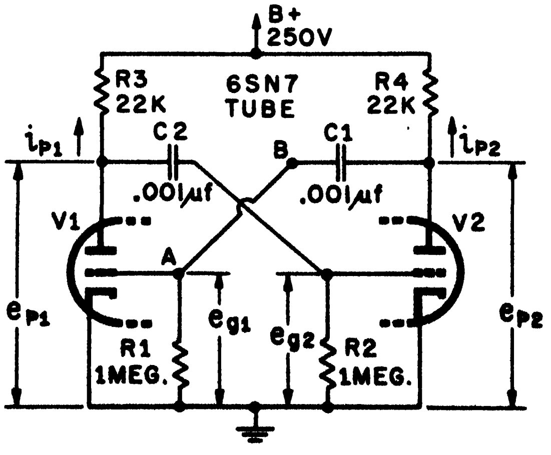


Fig. 9-51.—Circuit of a symmetrical plate-coupled multivibrator, giving typical component values.

sequently, we see that the feedback is regenerative, and if strong enough, oscillations result. The term plate-coupled multivibrator arises from the convention of the plate circuit of one tube being coupled to the grid circuit of the other tube. Typical values for a symmetrical type of plate-coupled multivibrator operating at a low frequency are those indicated in the drawing of Fig. 9-51. Let us examine this latter circuit and see how it operates as an oscillator.

With B+ supply applied to the tubes, the cathodes heated, and with initially zero bias, plate and grid current begin to flow in each tube. The initial grid-current flow is quite high and the grid-to-cathode resistance is considerably lower than that of the grid resistors. Consequently, as the plate voltages on the tubes increase, the capacitors in the circuit begin to charge up through their respective plate-load resistors.

Due to the symmetry of the circuit, the plate currents that initially start to flow in each tube are at first considered equal to each other. However, it is impossible for the currents to remain the same; because of some small circuit dissimilarity always being present, a slight difference between the two currents will always exist. It is this slight difference which starts the circuit acting as a multivibrator. Once the difference in current manifests itself, the circuit functions in such a manner as to increase this unbalancing effect. The plate and grid voltages and plate currents are indicated in the drawing of Fig. 9-51 by symbols. The circuit functions as follows:

Let us assume that the initial difference in current is such that the plate current i_{p1} of tube *V1* becomes slightly greater than plate current i_{p2} of tube *V2*. This increase in i_{p1} causes the voltage drop across the load resistor *R3* to increase, thereby causing the voltage, e_{p1}, at the plate of *V1* to decrease. The series combination of capacitor *C2* and resistor *R2* are in parallel with tube *V1*. Thus, with the voltage across *V1* decreasing, the voltage across *R2* and *C2* also decreases, so that when e_{p1} decreases, we actually have a decrease in e_{g2}, the grid voltage of tube *V2*. With a decrease in e_{g2}, the plate current, i_{p2}, of *V2* will also decrease, causing a smaller voltage drop in the plate-load resistor *R4*, and hence a larger voltage, e_{p2}, at the plate of *V2*. This increase in e_{p2} causes an increase in e_{g1}, and hence an increase in i_{p1}.

With this latter increase in i_{p1}, we are back to the original unbalanced condition, except that now the difference in the plate currents is greater than before. The cycle of operation starts all over again, until grid voltage e_{g2} is decreased to the point where i_{p2} is reduced to zero and i_{p1} is increased to a maximum. At this point, the plate voltage e_{p1} is a minimum and that of e_{p2} a maximum. This action described is a continuous building-up regenerative system that resembles a very rapid switching action and can be considered as occurring in an almost infinitesimal amount of time. We are now at a condition where tube *V2* is cut off, and tube *V1* has maximum conduction which can be assumed as the starting condition for the operation of the circuit.

The length of time that tube *V2* will be cut off and tube *V1* conducting will be the same, and a point will be reached where the actions are reversed, that is, *V1* will become cut off and *V2* conducting. This latter condition exists for the same amount of time as the former, and once again *V2* will become cut off and *V1* conducting. These actions are continuous and represent a rapid switching action where each tube is alternately conducting and cut off.

For any one tube to be cut off, the negative bias on the grid of the tube must be equal or beyond the cutoff voltage of the tube. The charge and discharge of the circuit capacitors play the important role in this bias operation. In order to better understand the function of the capacitors and how this switching action occurs, we will study the operation of the circuit in conjunction with its current and voltage waveforms.

Waveforms and Equivalent Circuits

The plate- and grid-voltage waveforms of the symmetrical plate-coupled multivibrator readily indicate the action within this circuit. In the following waveform analysis, in order to make the presentation clearer, equivalent circuits of the multivibrator of Fig. 9-51, during periods of capacitor charge and discharge, will be employed. In Fig. 9-52 are illustrated the waveshapes of the plate and grid voltages and currents of each tube of the multivibrator circuit. The equivalent multivibrator circuits for the conditions of charge and discharge of each capacitor are shown in Fig. 9-53. In the following discussion, continuous correlation should be made between the waveshapes of Fig. 9-52 and the circuits of Figs. 9-51 and 9-53.

For our starting conditions, we are assuming that tube *V1* has maximum conduction and *V2* has just been cut off. This is indicated at time *1* in curves (A) and (E) respectively of Fig. 9-52. At this same time, the plate voltage of *V1* is, therefore, a minimum and that of *V2* is rising toward a maximum, as indicated in curves (B) and (F) respectively. Since there is no longer a voltage drop across *R4* due to the plate current of *V2*, capacitor *C1* charges to the *B+* voltage. The rapid increase in plate voltage of *V2*, and hence the

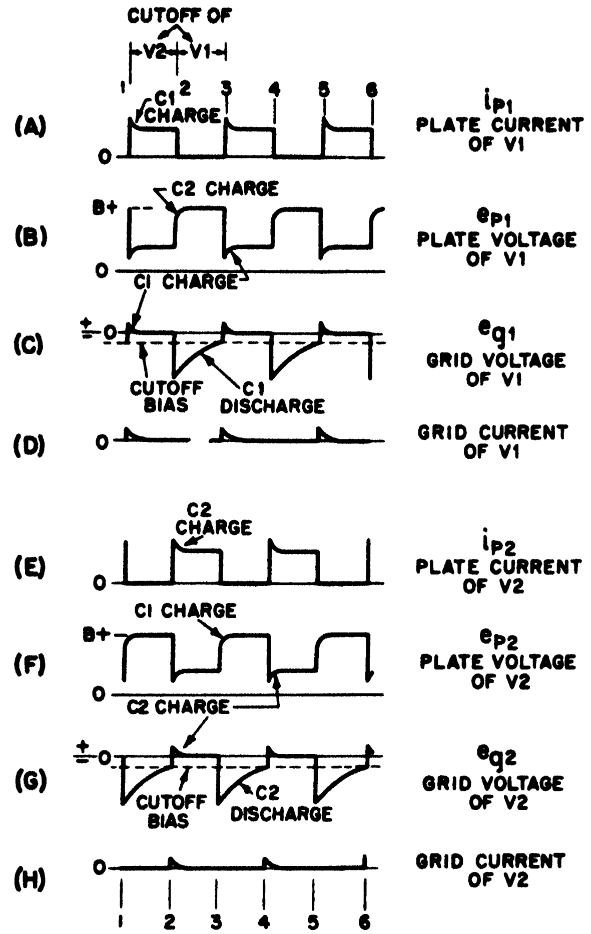


Fig. 9-52.—Drawings (A) through (H) illustrate the waveshapes of the plate and grid voltages and currents of each tube of the symmetrical multivibrator shown in Fig. 9-51.

charging voltage on *C1*, drives the grid positive and causes grid current to flow; this is shown in curves (C) and (D), respectively, of Fig. 9-52.

The charging path of *C1* is through *R4* and the grid-to-cathode resistance, r_{g1} of tube *V1*. The equivalent circuit is shown in Fig. 9-53A. Grid resistor *R1* is actually in parallel with r_{g1}, but the value of *R1* is very large compared with that of r_{g1}, so the former may be neglected and is, therefore, shown dotted. Although the charging voltage is quite high, the positive value of the grid voltage e_{g1} is limited because of grid-current flow through *R1*. The grid, driven positive, draws current which flows through *R1* and develops a negative voltage drop across this resistor; this opposes the positive voltage due to the charge on *C1*. After *C1* has become fully charged, plate current i_{p1} levels off to its maximum value and plate voltage e_{p1} levels off to its minimum value and the grid voltage falls to zero. Note, from curves (C) and (D) of Fig. 9-52, that the flow of grid current reduces to zero at the same rate as the grid voltage reduces to zero. The pips at time *1* of curves (A), (B), (C), and (D) of Fig. 9-52, as well as the bend in the plate-voltage curve (F), indicate the initial charge of capacitor *C1*.

At the *same instant*, point *1*, that capacitor *C1* starts to charge, capacitor *C2*, which is already fully charged, starts to discharge through the low plate resistance r_{p1} of tube *V1*, and resistor *R2*. The equivalent circuit during the discharge of capacitor *C2* appears in Fig. 9-53B. The initial discharge current is a maximum and causes a maximum voltage drop across *R2*, biasing *V2* beyond cutoff as indicated at time *1* in curve (G) of Fig. 9-52. As capacitor *C2* discharges, the current flowing through, and hence the voltage across *R2*, decreases exponentially until the grid voltage of *V2* reaches cutoff, as indicated at time *2* in curve (G) of Fig. 9-52, and *V2* starts conducting.

At this same time, the instantaneous switching action, described previously, occurs. However, in this case, the grid voltage and plate current of tube *V2* rise to their maximum values, as shown by curves (G) and (E), respectively, of Fig. 9-52, and the plate-voltage curve (F) decreases to a minimum. The functions of capacitors *C1* and *C2* reverse; that is, *C1* starts discharging and *C2* starts charging.

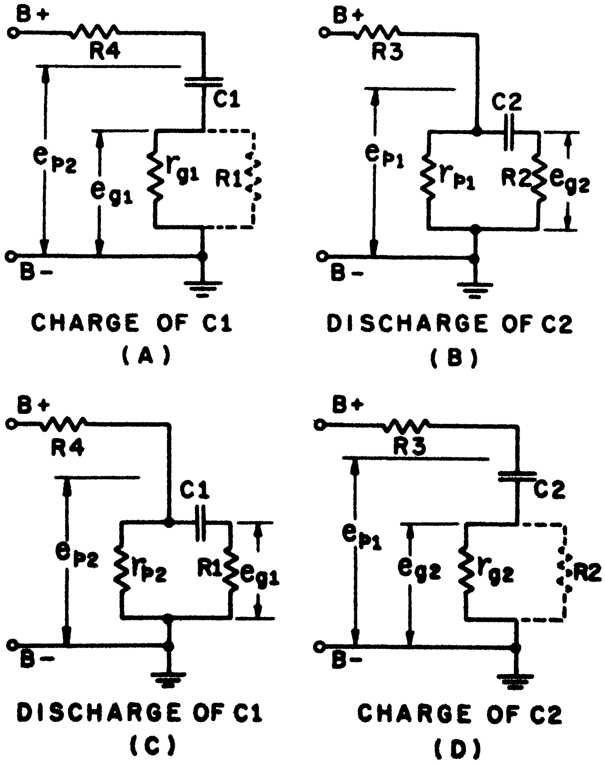


Fig. 9-53.—Equivalent circuits for the charging and discharging of capacitors *C1* and *C2* of the symmetrical multivibrator.

Capacitor *C1* discharges through *R1* and the plate resistance r_{p2} of conducting tube *V2*, as seen from the equivalent circuit of Fig. 9-53C. The initial discharge current of *C1*, which is a maximum, causes a voltage drop across *R1* sufficient to drive the grid of *V1* well beyond its cutoff point. This is indicated at time *2* of curve (C) of Fig. 9-52. This means that, at the same time, the plate current of *V1* is reduced to zero, shown in curve (A), *V1* will become nonconductive, and the plate voltage, therefore, increases to a maximum as seen in curve (B).

When the plate current reaches zero, capacitor *C2* charges to the value of the supply voltage. This charging voltage on *C2* drives the grid of *V2* slightly positive, so that it draws current, as seen at time *2* in curves (G) and (H) of Fig. 9-52. The charging path is through resistor *R3* and the grid-to-cathode resistance r_{g2} of tube *V2*. The equivalent circuit is shown in Fig. 9-53D. Grid resistor *R2*, being in parallel with r_{g2} and actually part of the charging circuit, is comparatively high, so that it can be neglected and is shown dotted in the equivalent circuit. The charging current of *C2* also prevents plate voltage e_{p1} from reaching its maximum value until the capacitor is fully charged and results in "pips" on the plate-current, plate-voltage, and grid-voltage waveforms at time *2*; this is indicated in curves (E), (F), and (G), respectively, of Fig. 9-52.

Capacitor *C1* continues to discharge at a decreasing exponential rate until the grid voltage e_{g1} of tube *V1* reaches cutoff. This occurs at time *3* on curve (C). At this instant, the switching process reverses once again, with tube *V1* becoming conductive and tube *V2* cut off. Capacitor *C1* starts charging and *C2* discharging. Tube *V2* is cut off by the initial discharge of capacitor *C2*. Thus, a complete cycle of multivibrator operation has occurred, the entire process starts over again, and a continuous off-on switching action between the tubes results.

For comparison purposes, we have taken oscilloscope pictures of the plate and grid voltages of the plate-coupled *symmetrical* multivibrator circuit of Fig. 9-51. A symmetrical or *balanced* type of multivibrator is one in which the circuit is completely balanced; that is, the time constants, tubes used, and applied voltages are the same such that the conducting and nonconducting time of each tube are the same. These oscillograms are illustrated in Fig. 9-54. The waveshapes of parts (A) to (D) are aligned one under the other, so that their cycles of operation can be compared. Note how the plate- and grid-voltage oscillograms of tube *V1*, parts (A) and (B), compare with the drawings of parts (B) and (C) of Fig. 9-52. Also note the point where the grid voltage of Fig. 9-54B has reached its cutoff bias value. The waveshapes pictured in parts (C) and (D) of Fig. 9-54 are the plate and grid voltages, respectively, of tube *V2*. Compare these latter two photos with the drawings of Fig. 9-52F and G, respectively. An interesting feature about the waveforms of both Figs. 9-52 and 9-54 that you may have noticed is that the plate and also the grid waveforms of each tube are similar in shape but are 180° out of phase with each other. If you were to shift the plate or grid voltage of either tube one half-cycle or 180°, these waveforms would then be exactly the same as those of the other tube.

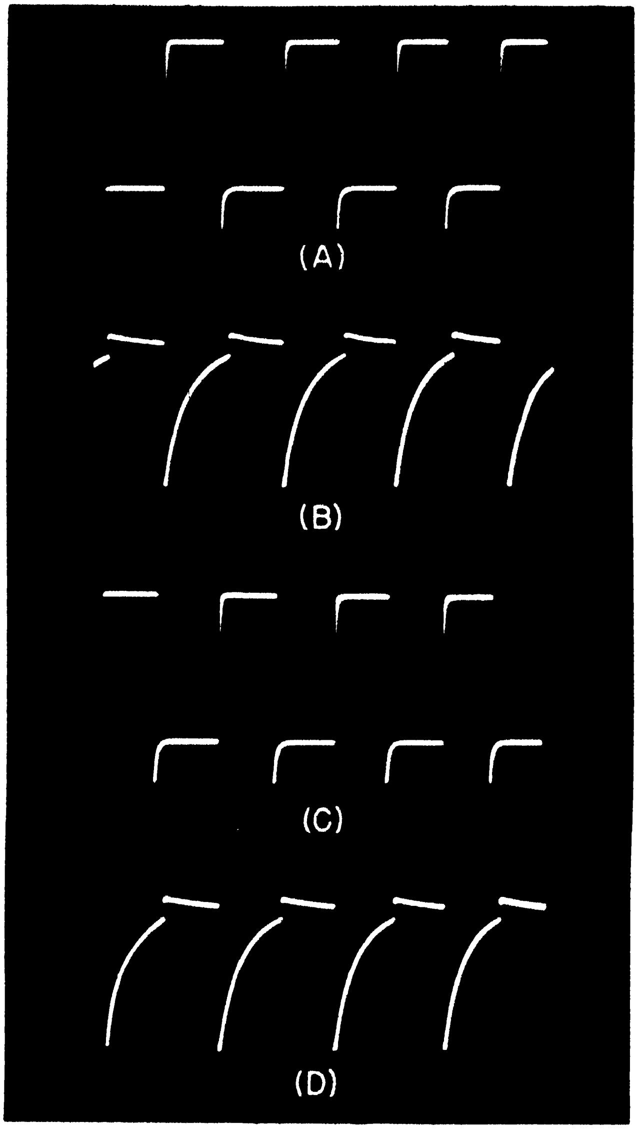


Fig. 9-54.—Oscillograms of the plate and grid voltages of the symmetrical multivibrator circuit of Fig. 9-51.

In order to see more clearly the charging and discharging portions of some of these waveforms of Fig. 9-54, one cycle of each has been expanded on the oscil-

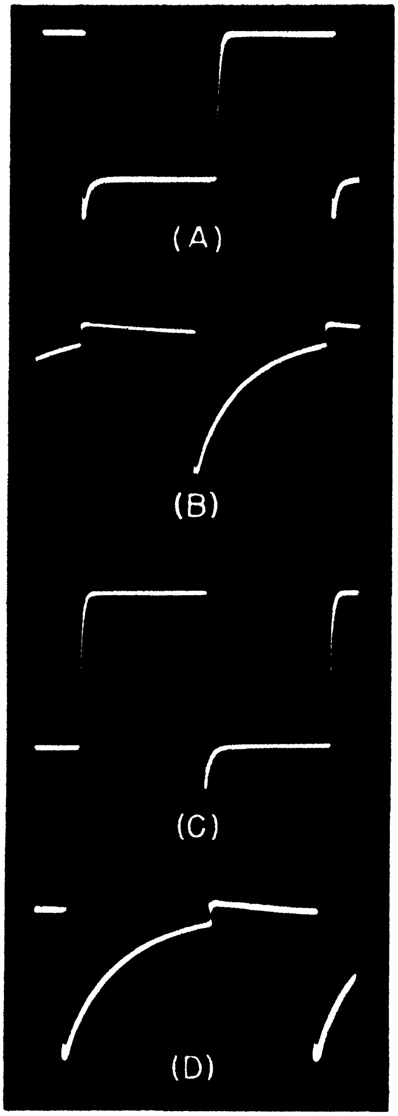


Fig. 9-55.—One cycle of each of the oscillograms shown in Fig. 9-54 has been expanded here to show more clearly the charge and discharge portions.

loscope and then photographed, the pictures of which appear in Fig. 9-55. Parts (A) to (D) of this latter figure correspond to parts (A) to (D) of Fig. 9-54.

Asymmetrical Multivibrator

If the *C1-R1* or *C2-R2* time constant in the circuit of Fig. 9-51 were changed, then the resulting waveforms would not be the same as those of Fig. 9-52. The time for completion of the waveshapes from points *1* to *2* of Fig. 9-52 would be different than that between points *2* and *3*. These unequal time constants would change the charging and discharging periods of the capacitors. This would result in one tube conducting for a longer period of time and the other tube conducting for a shorter period of time than for the previous symmetrical case where both the conduction and nonconduction times of each tube are always the same. In other words, the period of conduction for each tube would not be the same as its cutoff period. Such a circuit is known as an *asymmetrical* or *unbalanced* multivibrator.

If one of the time constants is made smaller, then the discharge time of that capacitor will also be smaller. Since the discharge of the capacitors in multivibrators controls the switching action of the circuit, then if the time constant of *R2-C2* of Fig. 9-51 is made shorter than that of *R1-C1*, tube *V2* will be cut off for a much shorter period of time than previously. This means that the resulting waveforms of Fig. 9-52 will be such that the first half-cycles, between points *1* and *2*, will have a shorter period. However, the second half-cycle, between points *2* and *3*, will remain substantially the same, since there has been no change in the discharge path of *C1*. If the time constant of *R1-C1* is shorter than that of *R2-C2*, the reverse will be true, and *V1* will be cut off for a shorter period of time. It is obvious, then, that changing either *R-C* time constant will change the frequency of the multivibrator.

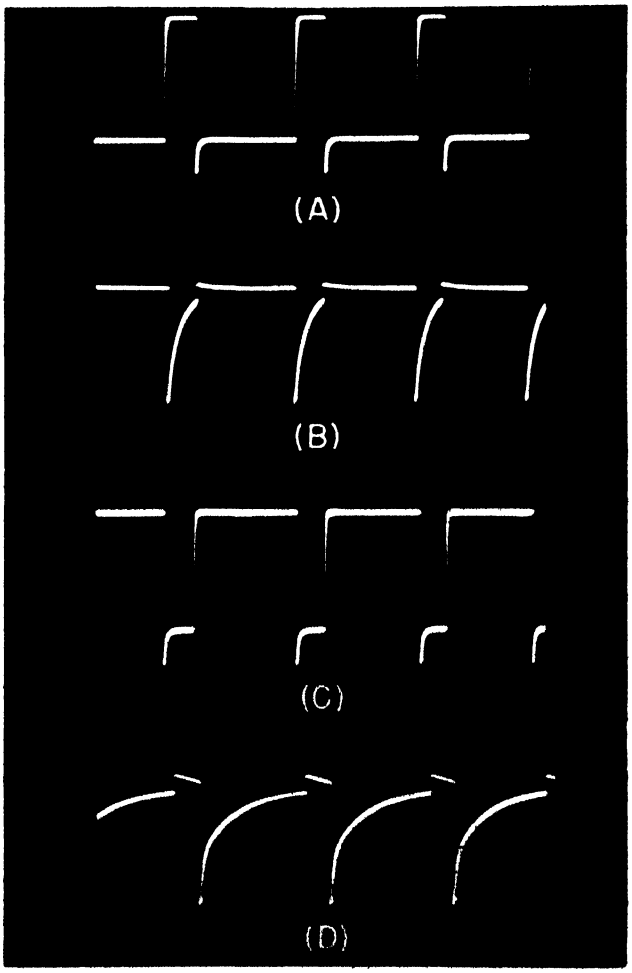


Fig. 9-56.—Plate- and grid-voltage oscillograms of the unbalanced multivibrator.

In order to indicate the waveforms under such conditions, we changed the *R1-C1* time constant of the symmetrical multivibrator circuit of Fig. 9-51 so that the circuit became asymmetrical. Resistor *R1* was changed to about 500,000 ohms and capacitor *C1* reduced to 700 $\mu\mu f$. Thus the time constant was changed from 0.001 second, or 1,000 microseconds, (1 megohm × 0.001 μf) to 350 microseconds (0.5 megohm × 0.0007 μf).

A number of oscilloscope pictures of the resulting patterns of this asymmetrical or unbalanced multivibrator are illustrated in Fig. 9-56. The waveforms of part (A) and (C) indicate the plate voltages of tubes *V1* and *V2* respectively. The grid voltages of these tubes under the same condition of a lower *R1-C1* time constant are shown in parts (B) and (D), respectively. Compare the photos of Fig. 9-54 with those of Fig. 9-56. The positive peaks of plate voltage of photo (A) indicate the period of plate-current cutoff of tube *V1*. Thus *V1* is not conducting for a shorter period of time than it conducts. The period of *C1* discharge is smaller, too, as indicated by the exponential rising characteristic of the grid-voltage waveform in part (B) of Fig. 9-56. Since *V1* is now conducting for a relatively longer period of time, tube *V2* is cut off for that same period of time. The longer time constant of *R2-C2*, as compared to that of *R1-C1*, is illustrated by the exponential discharge curve of *C2* in part (D) having a greater period of discharge than that of *C1* in part (B).

If the time constant of *R1-C1* is made still smaller, tube *V1* will be nonconductive for even a shorter period of time. If the time constants of the circuits were interchanged, that is, if *R2-C2* were made smaller than *R1-C1*, then tube *V2* would be nonconductive for the shorter period of time. Under this latter condition, the waveforms of (A) and (B) in Fig. 9-56 would represent those of tube *V2*, and those of (C) and (D) would be for tube *V1*.

Sawtooth Wave Output

In order to produce a sawtooth voltage waveform from a multivibrator circuit, periodic current pulses of short time duration are necessary. In this respect, the symmetrical multivibrator circuit cannot be used because the output waveforms from the plate circuits of both tubes have their positive and negative half-cycles equal to each other. If the time constant of one tube is made smaller than the time constant of the other tube, we have the asymmetrical multivibrator, and pulses of short time duration are then possible with this circuit. According to the photos of Fig. 9-56A and C, the plate voltage of tube *V1* has its positive pulses of short time duration and that of tube *V2* of long time duration. This means that the plate current of tube *V1* flows for a longer period of time, and that of tube *V2* for a shorter period of time or, as is commonly referred to, flows in pulses. The important question at the moment is: how do these current pulses help produce a sawtooth wave output?

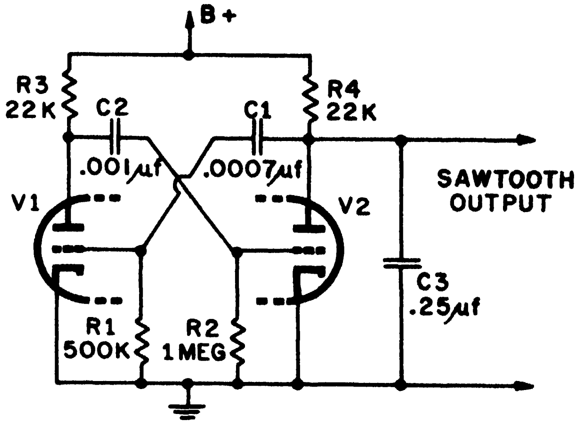


Fig. 9-57.—Circuit diagram of a plate-coupled asymmetrical (unbalanced) multivibrator.

In Fig. 9-57 we see the circuit of a plate-coupled asymmetrical multivibrator similar to that used in producing the waveshapes of Fig. 9-56. The only difference in the circuit of Fig. 9-57 is in the addition of a 0.25-μf capacitor, *C3*, across the plate-to-cathode circuit of *V2*. It is this capacitor, working in conjunction with tube *V2*, that produces a sawtooth wave output. The action is based upon the charging of capacitor *C3* to produce the rising part or trace of the sawtooth wave, and the discharge of *C3* to produce the retrace of the sawtooth. The over-all action is very similar to that of the thyratron sweep generator of Fig. 9-35. The sawtooth-producing action of the circuit of Fig. 9-57 is as follows:

Capacitor *C3* charges up from the power supply through resistor *R4*. The charge on this capacitor continues to build up during the nonconducting period of tube *V2*. As soon as tube *V2* starts to conduct, it represents a very low resistance across *C3*. This low resistance is equivalent to a short across *C3* and the capacitor begins to discharge through *V2*. Since the conduction period of *V2* is short (i.e. its plate current flows in pulses of short time duration), the period of discharge will be very rapid because this effective short-circuit across *C3* only lasts for the time that *V2* conducts.

A picture of the sawtooth wave output is illustrated in Fig. 9-58. If the time duration of the plate-current

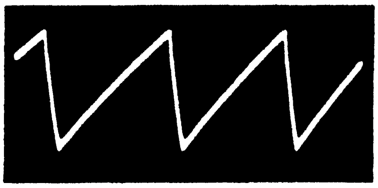


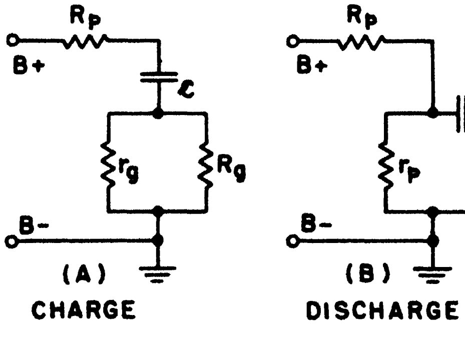
Fig. 9-58. — Oscillogram of the sawtooth-wave output produced by the unbalanced multivibrator shown in Fig. 9-57.

pulses of *V2* were decreased still further, the resulting sawtooth output would have a flyback time shorter than that illustrated. The time constant of the *C3-R4* charging circuit determines the slope of the charging curve. In order to assure a fairly linear sawtooth output, the value of plate resistor *R4* is usually made higher than that indicated.

Time Constants and Frequency of Oscillation

The frequency of oscillation of the multivibrator circuit is primarily dependent upon the two time constants of the circuit. This is true even when pentodes are used instead of triodes. However, there are slight differences in each case which will soon be seen. Let us first consider the symmetrical triode multivibrator circuit of Fig. 9-51 and the equivalent circuits, under conditions of charge and discharge, as shown in Fig. 9-53. In a symmetrical multivibrator, all circuit components for each tube are usually made equal to each other, including the tube characteristics and applied voltages. Under this condition, the equivalent charging circuits of Fig. 9-53 (A) and (D) are the same, as are the discharge circuits of (B) and (C). In the final analysis, then, for a symmetrical multivibrator we can consider only one equivalent circuit for charging and one for discharging. This is indicated in the redrawn equivalent circuits as shown in Fig. 9-59A and B. In these circuits, R_p is the plate-load resistor, R_g is the grid resistor, r_p the plate resistance of the tube, and r_g the grid-to-cathode resistance during grid-current flow.

Many texts refer to the time constants of either tube circuit of a multivibrator as being approximately equal to the product of the capacitor and associated grid resistor. However, to be more accurate, we will consider the time constants as presented by the equivalent circuits of Fig. 9-53 and draw our approximations later for triodes and pentodes.



(A) CHARGE (B) DISCHARGE

Fig. 9-59.—Basic equivalent circuits for the charge and discharge of a symmetrical multivibrator.

The time constant during periods of charge T_c for a symmetrical multivibrator circuit is, as seen from Fig. 9-59A, equal to $C \times R_t$, where R_t is the total or equivalent resistance of the circuit. In the circuit under discussion, R_t is equal to the sum of R_p and the parallel resistance of R_g and r_g. Expressed mathematically

$$T_c = C\left(R_p + \frac{R_g r_g}{R_g + r_g}\right). \tag{9-2}$$

During periods of discharge, the time constant T_d has a different equivalent resistance, which is equal to the sum of grid resistor R_g and the parallel resistance of R_p and r_p. (These two latter resistances are effectively in parallel through the low resistance of the B supply.) The time constant during periods of discharge is, therefore

$$T_d = C\left(R_g + \frac{R_p r_p}{R_p + r_p}\right). \tag{9-3}$$

If you examine the waveforms of the symmetrical multivibrator of Fig. 9-52, you will note those portions of the curves that are labelled "*C1* charge" and "*C2* charge" actually do not occupy a complete half-cycle of waveform. On the other hand, the discharge periods of each of these capacitors are the same as one half-cycle of the waveform. This is readily seen in curve (G) of Fig. 9-52, which indicates that during discharge of *C2*, the first half-cycle of grid voltage of tube *V2* is completed; during discharge of *C1*, seen in part (C), the second half-cycle of grid-voltage waveform of *V1* is completed. Thus the discharge periods of *C1* and *C2* represent a complete cycle of operation. From these drawings, we can state that the frequency of oscillation of the multivibrator is primarily dependent upon the discharge time constants of both capacitors. This statement is not completely accurate, since it neglects the charging portion of the curves, but for most approximations this may be disregarded. In order to use the time constant of equation (9-3) in computing the frequency of oscillation, the plate and grid voltages have to be taken into account.

In dealing with a symmetrical multivibrator circuit, we can use the same symbols of Fig. 9-51, without numerical subscripts, in figuring the frequency of operation. Thus e_g is the grid voltage, e_p the plate voltage, and *C* the capacitor for each tube. Whatever is said about one discharge circuit also holds true for the other. The discharge-current curve of the capacitor decreases exponentially. Therefore, as capacitor *C*·

discharges, the grid voltage e_g across the grid resistor also decreases exponentially and is represented by the equation

$$e_g = E\epsilon^{-\frac{t}{R_tC}} \qquad (9\text{-}4)$$

where t represents time, R_t the total equivalent resistance in the discharge circuit, C the discharge capacitor, and E the *negative* voltage to which the grid is being driven. E is equal to the supply voltage E_b, less the equilibrium value of tube plate voltage e_p. (This difference, $E_b - e_p$, is the same as the voltage drop across either plate-load resistor since these resistors are equal.) Thus equation (9-4) can now be written

$$e_g = (E_b - e_p)\epsilon^{-\frac{t}{R_tC}}. \qquad (9\text{-}5)$$

We can make use of equation (9-5) to obtain an expression for the frequency of operation for the circuit of Fig. 9-51. Let us first rearrange the equation by dividing both sides by e_g. Therefore, we have

$$1 = \left(\frac{E_b - e_p}{e_g}\right)\epsilon^{-\frac{t}{R_tC}}. \qquad (9\text{-}6)$$

Now, taking the natural logarithm of both sides of equation (9-6), we find[6]

$$\log_\epsilon 1 = \log_\epsilon\left(\frac{E_b - e_p}{e_g}\right) + \log_\epsilon \epsilon^{-\frac{t}{R_tC}} \qquad (9\text{-}7)$$

After simplifying, we have

$$0 = \log_\epsilon\left(\frac{E_b - e_p}{e_g}\right) - \frac{t}{R_tC}. \qquad (9\text{-}8)$$

Rearranging equation (9-8), and solving for time t, the result is

$$t = R_tC \log_\epsilon\left(\frac{E_b - e_p}{e_g}\right). \qquad (9\text{-}9)$$

Equation (9-9) tells us the time it will take either capacitor of a symmetrical plate-coupled multivibrator to discharge. Since this equation holds for either capacitor, then twice the value of this equation will be the time it will take to complete one cycle of operation. Therefore, the total period of time is

$$t = 2R_tC \log_\epsilon\left(\frac{E_b - e_p}{e_g}\right). \qquad (9\text{-}10)$$

The reciprocal of this complete time $1/t$ is, therefore, equal to the frequency of operation. Thus

$$f = \frac{1}{2R_tC \log_\epsilon\left(\frac{E_b - e_p}{e_g}\right)}. \qquad (9\text{-}11)$$

Equation (9-11) is the expression for the frequency of operation of a symmetrical plate-coupled multivibrator circuit. Substituting the equivalent resistance of discharge from equation (9-3) for R_t in equation (9-11) we have

$$f = \frac{1}{2C\left(R_g + \frac{R_pr_p}{R_p + r_p}\right) \log_\epsilon\left(\frac{E_b - e_p}{e_g}\right)}. \qquad (9\text{-}12)$$

Equation (9-12) will apply to a plate-coupled multivibrator circuit using pentode tubes, especially when the plate resistances of the tubes and the plate-load resistors are comparable in value to the grid resistors. For triodes, the plate resistance of the tube is usually quite small compared to the plate-load resistor, so that the parallel resistance of R_p and r_p is approximately equal to r_p alone. Compared to the grid resistor R_g usually employed, the value of r_p for a triode is quite small, so that equation (9-12) for triodes, employed in symmetrical plate-coupled multivibrators, can for most approximations be written

$$f = \frac{1}{2R_gC \log_\epsilon\left(\frac{E_b - e_p}{e_g}\right)}. \qquad (9\text{-}13)$$

For unsymmetrical plate-coupled multivibrators, the frequency of operation is different because the values of grid resistors, discharging capacitors, plate voltages, and grid voltages are no longer equal to each other. For unbalanced multivibrators, we also have to deal with two time equations similar to equation (9-9) because the discharge times of the two capacitors are no longer equal. For capacitor *C1* of Fig. 9-51, we have to deal with the grid voltage e_{g1} on tube *V1*, the plate voltage e_{p2} on tube *V2*, and a different equivalent discharge resistance for each tube. Then for capacitor *C1*, equation (9-9) for asymmetrical multivibrators becomes

$$t_1 = R'_tC_1 \log_\epsilon\left(\frac{E_b - e_{p2}}{e_{g1}}\right) \qquad (9\text{-}14)$$

and for capacitor *C2*, equation (9-9) becomes

$$t_2 = R''_tC_2 \log_\epsilon\left(\frac{E_b - e_{p1}}{e_{g2}}\right). \qquad (9\text{-}15)$$

In these last two equations, R'_t is the equivalent resistance in the discharge path of *C1*, and R''_t the equivalent resistance in the discharge path of *C2*, as indicated in parts (C) and (B) of Fig. 9-53. Therefore

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[6] The natural logarithm of 1 is zero and the natural logarithm of ϵ is equal to 1. If ϵ is raised to some exponent, as in equation (9-7), then the natural logarithm of the complete expression is equal to the exponent, including the sign in front of the exponent.

$$R'_t = R1 + \frac{(R4)(r_{p2})}{(R4) + (r_{p2})} \qquad (9\text{-}16)$$

and

$$R''_t = R2 + \frac{(R3)(r_{p1})}{(R3) + (r_{p1})}. \qquad (9\text{-}17)$$

The frequency of operation of an asymmetrical plate-coupled multivibrator can be then written as

$$f = \frac{1}{t_1 + t_2}. \qquad (9\text{-}18)$$

Substituting equations (9-14) and (9-15) in equation (9-18), we have

$$f = \frac{1}{R'_tC_1 \log_{\epsilon}\left(\frac{E_b - e_{p2}}{e_{g1}}\right) + R''_tC_2 \log_{\epsilon}\left(\frac{E_b - e_{p1}}{e_{g2}}\right)}. \qquad (9\text{-}19)$$

Equation 18 will be true for unbalanced plate-coupled multivibrators when pentode tubes are employed. Equations 15 and 16 can be substituted for R'_t and R''_t, respectively in equation 18 for a more complete expression for the frequency of operation.

When triode tubes are employed in the asymmetrical circuit, the equivalent resistances of equations (9-16) and (9-17) can, for most approximations, be simply equal to resistors *R1* and *R2*, respectively. Under this condition, equation (9-19) reduces to

$$f = \frac{1}{R_1C_1 \log_{\epsilon}\left(\frac{E_b - e_{p2}}{e_{g1}}\right) + R_2C_2 \log_{\epsilon}\left(\frac{E_b - e_{p1}}{e_{g2}}\right)}. \qquad (9\text{-}20)$$

If the expression $\left(\log_{\epsilon}\frac{E_b - e_p}{e_g}\right)$ in equations (9-11) and (9-13) is approximately equal to unity, the general equation (9-11) for symmetrical plate-coupled multivibrators becomes

$$f \approx \frac{1}{2R_tC} \qquad (9\text{-}21)$$

and, for triodes, equation (9-13) becomes

$$f \approx \frac{1}{2R_gC}. \qquad (9\text{-}22)$$

In like manner, if both natural logarithms of the voltage expressions in equations (9-19) and (9-20) are approximately equal to unity, then equation (9-19), the general expression for the frequency of an unbalanced plate-coupled multivibrator, becomes

$$f \approx \frac{1}{R'_tC_1 + R''_tC_2} \qquad (9\text{-}23)$$

and, for triodes, equation (9-20) becomes

$$f \approx \frac{1}{R_1C_1 + R_2C_2}. \qquad (9\text{-}24)$$

Cathode-Coupled Multivibrators

Another very common type of multivibrator, called the cathode-coupled multivibrator, is shown in Fig. 9-60. It obtains its name by virtue of the common cathode resistor R_k. This multivibrator circuit is used to produce different types of pulse waveforms among which the square wave is common. However, as in the plate-coupled type, this circuit can also be made to produce a sawtooth voltage output.

The circuit of Fig. 9-60 is of the free-running type. Most of the symbols used in this drawing are similar to that of the plate-coupled multivibrator of Fig. 9-51. Let us examine its operation. When the tubes are heated and plate voltage is applied to the circuit, both tubes conduct and capacitor *C1* charges to the supply potential. The plate current of both tubes flows through the common cathode resistor, making the cathodes more positive than ground, thereby causing a bias to be effectively applied to both grids, which reduces the current through each tube. Let us assume that due to some momentary unbalance in the circuit, the plate current of *V2* is slightly higher than the plate current of tube *V1*. The increased plate current i_{p2} causes a reduction in the plate voltage of tube *V2* due to the increased voltage drop across *R4*. This change in voltage is coupled to the grid of *V1* via *C1*, lowering the grid voltage and hence the plate current of *V1*. This reduced i_{p1} current causes a smaller voltage drop across R_k and hence a less negative bias across *V2*. Thus the plate current of *V2* is increased further and the plate voltage e_{p2} is decreased still more. This operation, which is considered to be instantaneous, continues until tube *V1* is cut off and tube *V2* is conducting at a maximum.

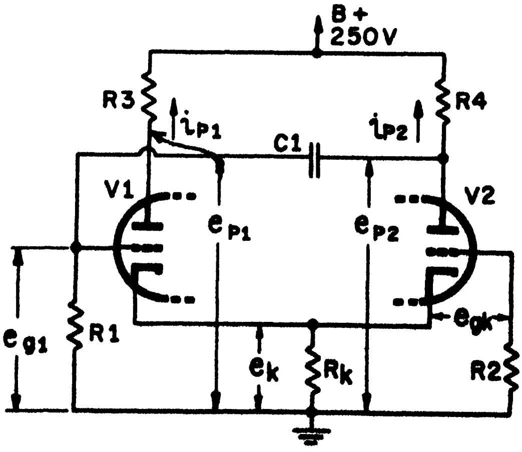

Fig. 9-60.—Circuit diagram of the cathode-coupled multivibrator.

Starting from this point of one tube being conductive and the other nonconductive, we can easily analyze the oscillator action of this circuit in conjunction with some waveforms of operation. Thus the plate voltage e_{p1} of tube *V1* is a maximum (*V1* is not conducting and i_{p1} is equal to zero) and the plate voltage e_{p2} of tube *V2* virtually a minimum (*V2* is conducting and i_{p2} is a maximum). At this starting point, voltage e_{g1} at the grid of *V1* is a maximum negative, and the cathode voltage e_k (across R_k) is approximately a minimum.

What actually happens is that when *V2* becomes conductive and *V1* is cut off, capacitor *C1* starts to discharge through resistors *R1* and R_k, and the plate resistance of tube *V2*. At the start of discharge, the discharging current is a maximum and a maximum negative voltage, well beyond the cutoff bias value of *V1*, exists acroo *R1* and hence at the grid of *V1*. The discharging current decreases exponentially, and as soon as the grid cutoff voltage of *V1* is reached, *V1* will start conducting.

Examine the waveforms of Fig. 9-61. That at (A) is the grid voltage e_{g1} of tube *V1;* at (B) is the plate voltage e_{p1} of tube *V1;* and at (C) is the cathode voltage across resistor R_k. The effective grid bias at *V2*, e_{gk}, measured between grid and cathode, is the inverse of e_k and is shown at (D); at (E) is the plate voltage e_{p2} of tube *V2*. Point *1* indicates the time where e_{g1} is most negative, hence when *C1* has had its initial discharge. Since *V1* is then cut off, voltage e_{p1} is a maximum, and e_k is close to its minimum value. The bias on *V2* is determined only by the voltage drop across R_k; therefore, *V2* is conducting heavily, reducing e_{p2} to a low value.

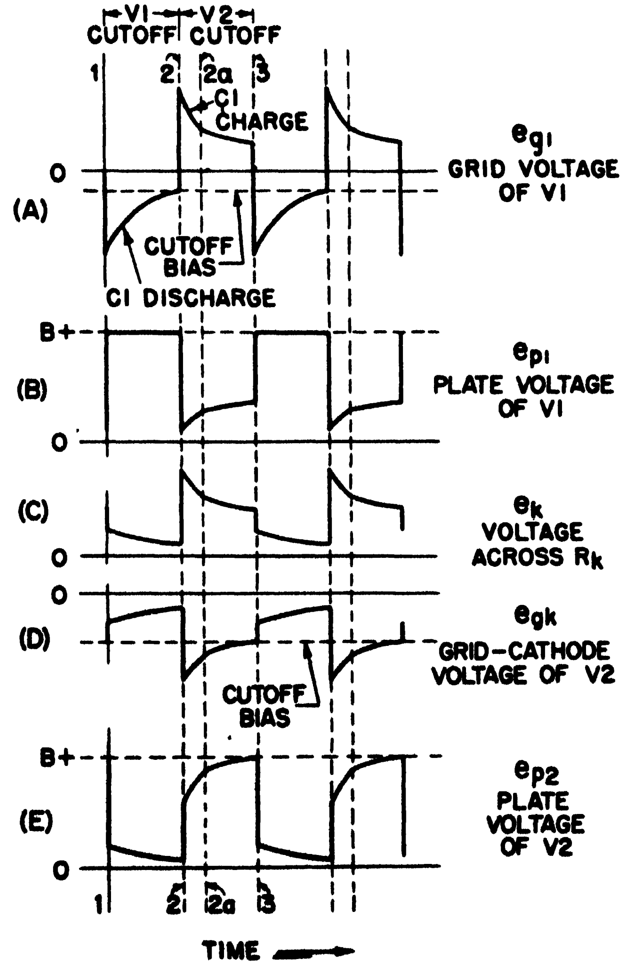

Fig. 9-61.—Plate-, grid-, and cathode-voltage diagrams of cathode-coupled multivibrator.

During the interval from *1* to *2*, the current through R_k is the plate current of *V2* plus the discharge current of *C1*. As the discharge current diminishes, the voltage across R_k decreases, lowering the bias e_{gk} of tube *V2*. However, the magnitude of the discharge current is small compared to that of the *V2* plate current, so that there is but a small change in the grid bias and plate voltage of *V2*, shown in curves (D) and (E) of Fig. 9-61. As *C1* discharges through *R1* and the low plate resistance of *V2*, the cutoff bias of *V1* will soon be reached. *V1* will start conducting, causing a greater voltage drop across R_k, which, in turn, will cause a decrease in *V2* plate current and an increase in e_{p2}. The increase in e_{p2} is coupled to the grid of *V1* through *C1*, driving the grid positive, as indicated at time *2* of curve (A). The plate current of *V1* increases further, and hence the voltage drop across R_k also increases. This instantaneous action is cumulative, reaching a point where voltage e_k is great enough to drive *V2* beyond cutoff, since e_{gk} is the reverse of e_k. Thus, at time *2* in Fig. 9-61, we find *V2* cut off and *V1* conducting.

It should be noted that a given tube cannot cut itself off by means of cathode bias. In this circuit, the two tubes are identical and are operated at the same plate voltage. Therefore, neither *V1* nor *V2* can ever cut each other off by virtue of cathode bias *alone*, when that bias is the result of plate current flowing through the cathode resistor and where the value of that plate current is determined *only* by the cathode bias. However, in the circuit of Fig. 9-60, the increasing plate voltage of *V2* is coupled to the grid of *V1* via capacitor *C1*, driving the grid positive by an amount which is sufficient to overcome the bias due to the voltage across R_k. Thus the plate current of *V1* is controlled primarily by the grid bias due to the charging current of *C1*, rather than by the cathode bias across R_k. Consequently i_{p1} increases, causing e_k to rise to a value which will drive *V2* beyond cutoff.

When the grid of *V1* is driven positive, grid current flows and capacitor *C1* charges through *R4*, and also

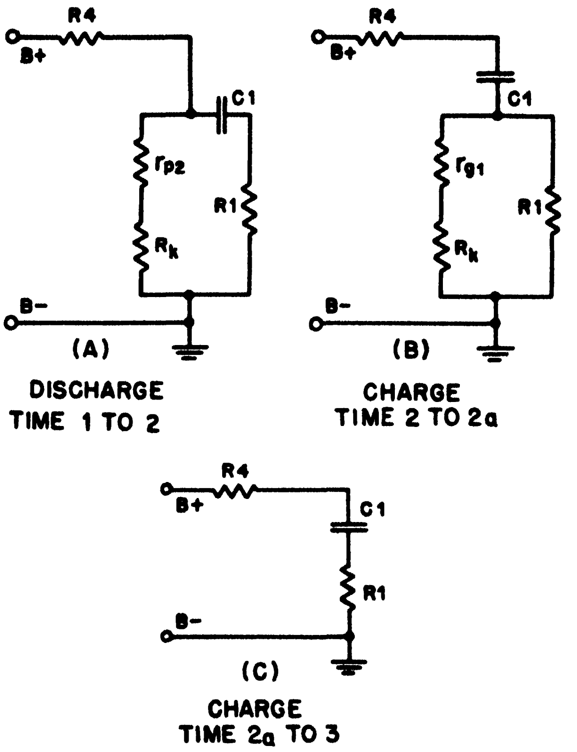

Fig. 9-62.—Equivalent circuits for charge and discharge of the cathode-coupled multivibrator.

through R_k and the low grid-to-cathode resistance of *V1,* both in parallel with *R1.* As *C1* charges, the positive bias on *V1* decreases exponentially. A point will be reached during the charging of *C1* where the positive bias on *V1* will be approximately equal to e_k, the voltage drop across R_k. This is indicated at time *2a* in Fig. 9-61. This means that grid current will stop flowing in *V1* because its grid-to-cathode voltage is now effectively zero.

The charging path after point *2a* is different because the low grid-to-cathode resistance of *V1* during grid-current flow no longer exists. The path is now through *R4* and *R1* only. This new time constant of charge after point *2a* is higher than that for the charge circuit between times *2* and *2a.* Thus *C1* continues to charge, but at a slower rate, as seen by the grid-voltage waveform of Fig. 9-61A. Remember that the grid of *V1,* after time *2a,* is positive only with respect to ground, but not with respect to its cathode. As e_{g1} decreases from point *2a,* i_{p1} also decreases, causing a reduction in the voltage across R_k. The plate voltage of *V2,* e_{p2}, does not rise to the supply voltage because of the voltage drop across *R4* due to the charging current of *C1;* this is shown in curve (E) of Fig. 9-61. Note that there is no coupling from the plate of tube *V1* to the grid of *V2* and only resistor *R2* exists between grid and ground of *V2.* Hence the grid is always at ground potential, and tube *V2* will remain cut off as long as the potential across R_k is greater than the cutoff bias of the tube.

Now a point will be reached where the reduced i_{p1} flowing through R_k will reduce e_k to the point where the effective bias on *V2* reaches cutoff; this occurs at time *3* in curves (C) and (D). When this point is reached, tube *V2* starts conducting, *C1* begins to discharge again through *V2,* and tube *V1* becomes cut off. *V1* is cut off because the high discharging current of *C1* flows through *R1,* biasing *V1* beyond cutoff. We have now completed one cycle of operation of this cathode-coupled multivibrator. After point *3* in Fig. 9-61, the complete cycle repeats itself.

We have previously explained that two separate charging paths have to be considered for the cathode-coupled multivibrator of Fig. 9-60, and only one discharging path. The three equivalent circuits for the charging and discharging of *C1* are illustrated in Fig. 9-62. The equivalent discharge circuit is shown at (A); it exists during time *1* to *2* of Fig. 9-61 (during the first half-cycle). At (B) is the charging path of *C1* during time *2* to *2a,* which occurs during the period of grid-current flow in *V1.* From time *2a* to *3,* grid current no longer flows in *V1;* hence the value of the grid-to-cathode resistance r_{g1} is considered to be infinite compared to the other resistances in the circuit. Consequently, r_{g1} and R_k in Fig. 9-62B are no longer effective in the circuit, and the new equivalent charging circuit after time *2a* reduces to that shown at (C).

Typical Circuits and Oscillographic Waveforms

Fig. 9-63 illustrates a typical cathode-coupled multivibrator circuit used today to produce a fairly good

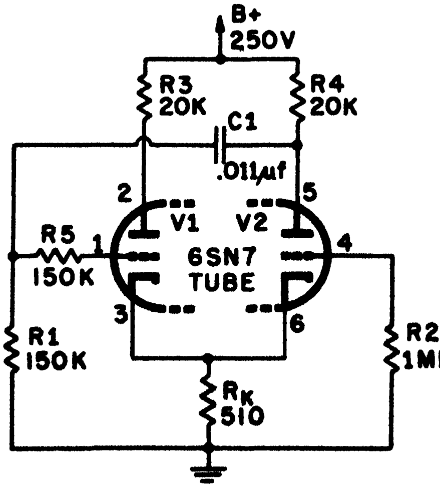

Fig. 9-63.—Typical commercial cathode-coupled multivibrator circuit for producing a good square-wave output waveform.

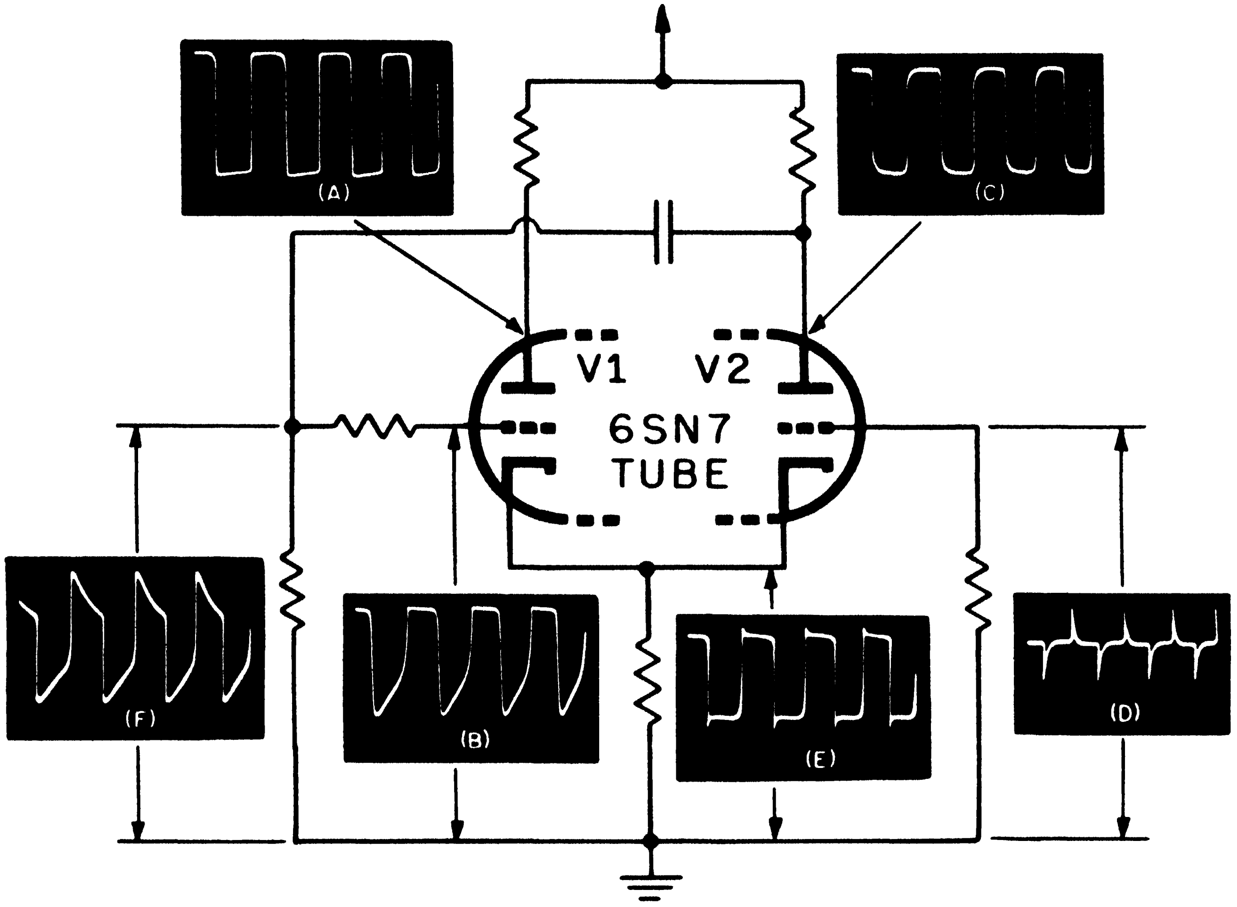

Fig. 9-64.—Simplified schematic of the circuit of Fig. 9-63 showing the various input and output waveforms.

square-wave output. The primary difference between this circuit and that of Fig. 9-60 is that here series resistor *R5* is inserted in the grid circuit of tube *V1*. The primary purpose of this resistor is to insure good flat-topped square-wave output. The resulting waveforms of operation are shown in Fig. 9-64 in conjunction with the circuit of Fig. 9-63, but with component numbers and values removed. Waveform (A) is the voltage from the plate of tube *V1* to ground, and is considered to be a fairly good square-wave output. Waveform (B) is from the grid of tube *V1* to ground. The square-wave output from the plate circuit of *V2* is indicated by waveform (C). Compare this waveform with that of (B) and note the better squaring of the former. Waveform (D) is taken between the grid of *V2* and ground. An attempt to obtain an oscillogram of the grid-to-cathode signal was unsuccessful, caused by stray pick-up, apparently due to the fact that the ground lead of the oscilloscope was not at the multivibrator circuit ground. The voltage waveform across the cathode resistor is illustrated by pattern (E). This signal is in the form of a square wave where the leading edge of the positive and negative flat tops (going from left to right) has a rising characteristic. Waveform (F) represents the signal existing between the junction of resistors *R1* and *R5* and ground.

For comparison purposes, the resistance of *R1* of Fig. 9-63 was changed to 22,000 ohms, and a number of waveforms were photographed. These oscillograms are shown in Fig. 9-65. That at (A) is the output from the plate of *V1*, (B) is the waveform between the grid of *V1* and ground, (C) is the signal at the plate of *V2*, and (D) the signal between the grid of *V2* and ground. The voltage across the cathode resistor is shown at (E), and (F) indicates the waveform that exists between ground and the junction of *R1* and *R5*. Comparison between the waveforms of Figs. 9-64 and 9-65 readily indicates that if the grid resistor *R1* is made too low, then the resulting waveforms lose their square-wave appearance. All of the oscillograms of Figs. 9-64 and 9-65 were photographed from an oscilloscope which employed return-trace blanking.

In the circuits of Figs. 9-60 and 9-63, the frequency of operation of the multivibrators is primarily depend-

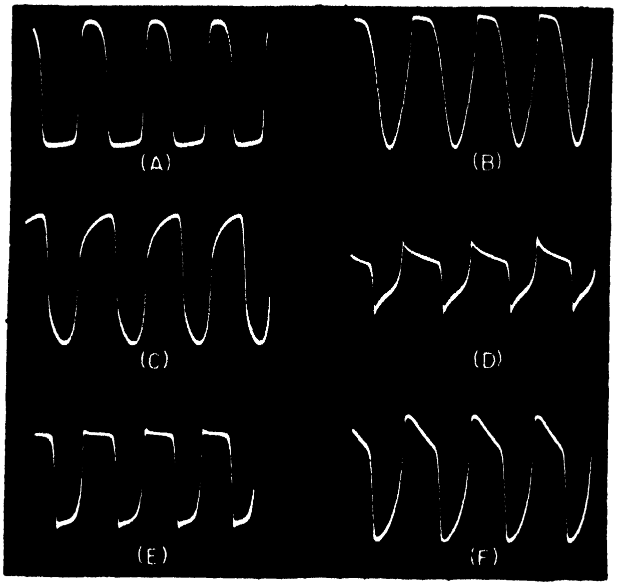

Fig. 9-65.—Oscillograms of waveforms produced by the multivibrator circuit of Fig. 9-63, after the resistor *R1* was changed to 22,000 ohms.

ent upon capacitor *C1* and resistor *R1*. Changing the value of either one of these components will control the frequency of operation of the circuit.

The cathode-coupled multivibrators illustrated thus far can be classified as the symmetrical type, because the periods of conduction and cutoff for each tube are the same and each cycle of the resulting waveforms is symmetrical in shape. In order to produce a good sawtooth output, the circuit must be unsymmetrical, similar to that of the unbalanced plate-coupled multivibrator. The primary means of accomplishing this unbalance with the cathode-coupled multivibrator is to employ different load resistors for each tube and/or to use different supply voltages. Both of these methods are commonly employed. Another possible method of establishing these asymmetrical qualities, although not used very much, is to employ a different type tube for each part of the multivibrator circuit. Besides these features, a capacitor is placed across the plate to ground of one of the tubes. This is similar to the method employed for the plate-coupled multivibrator of Fig. 9-57.

To obtain a sawtooth wave that has a short flyback time, the cutoff time of the tube across which the sawtooth capacitor is placed should be long compared to its period of conduction. This capacitor charges up from the power supply. When the tube in question starts conducting, it offers a low-impedance path to the capacitor which then begins to discharge through the tube. This low impedance exists only during tube conduction. The charge of the capacitor represents the linear portion of the sawtooth output and the period of discharge represents the flyback time. The shorter the conduction time of the tube is, the smaller is the period of capacitor discharge, and hence the less is the flyback time.

Fig. 9-66 illustrates a typical cathode-coupled multivibrator used to produce an output sawtooth waveform at a frequency of approximately 15 kc. Such circuits are often used as the linear sweep for certain television receivers. (15,750 cps is the horizontal sweep frequency used in television receivers.) When resistor *R4* is made variable, it is primarily used to regulate the amplitude of the sawtooth output signal. Remember that it is this resistor through which capacitor *C2* is charged. If resistor *R1* or capacitor *C1* is made variable, then the frequency of operation of the circuit can be controlled. In practice, *R1* is usually made the variable and in television receivers is often termed the "hold" control.

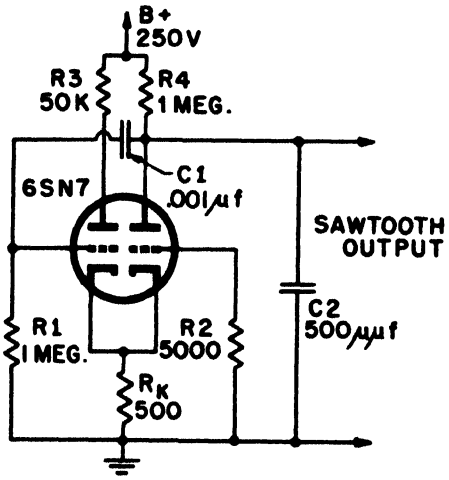

Fig. 9-66.—Cathode-coupled multivibrator circuit for producing a 15-kc output sawtooth waveshape.

The cathode-coupled multivibrator, when used to produce a sawtooth signal in the manner previously described, is often termed the "Potter's sweep circuit," after J. L. Potter.[7] Besides being used in television receivers, it also finds wide application as the sweep circuit for cathode-ray oscilloscopes. A typical cathode-coupled multivibrator circuit as used in oscilloscopes today is illustrated in Fig. 9-67. This circuit is very similar to that of Fig. 9-66 except that we have indicated those components which are usually made variable, with their respective values.

This circuit uses a 6J6 duo-triode tube. Resistor *R1* and capacitor *C1* are the primary frequency-determin-

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[7] Potter, J. L., "Sweep Circuit," *Proc. I.R.E.*, vol. 26, pp. 713-719, June 1938.

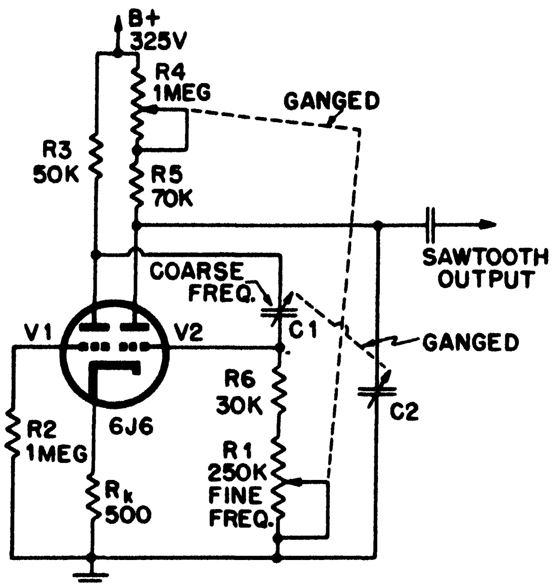

Fig. 9-67.—A typical cathode-coupled multivibrator circuit used in commercial oscilloscopes.

ing components and are made variable. The choice of *C1* is actually made by a switch, and not a rotating capacitor, and is often termed the "coarse frequency" control. You will note that capacitor *C2* is also variable and is ganged to *C1*. Capacitor *C2* is not a frequency-determining component; its chief function is to help shape the sawtooth signal output. As the coarse frequency is changed, this output capacitor will also be changed so that the shape of the output sawtooth signal will not vary with changes in frequency.

Also note that resistor *R4* is variable and is ganged to resistor *R1*. Both of these potentiometers have control over the frequency of operation; this control is, however, sharper than that of *C1* and is, therefore, often termed the "fine frequency" control or "sweep frequency vernier" control. Resistor *R4* regulates the voltage on the plate of *V2*, and, therefore, has a certain degree of frequency control, as well as amplitude control, of the resulting sawtooth output signal. The ganging between *R1* and *R4* is usually adjusted so that when the fine frequency control is varied, the output signal will change in frequency only. The circuit as it stands now will, with proper choice of *C1* from approximately 0.5 μf to 100 $\mu\mu f$, produce a sawtooth frequency output anywhere from approximately 5 cps to 60 kc.

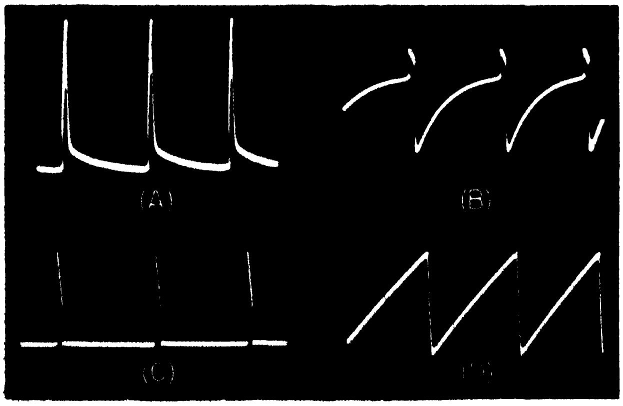

Fig. 9-68.—Oscillograms of waveforms for normal operating conditions of the commercial multivibrator circuit shown in Fig. 9-67.

In order to indicate the operating conditions in such a circuit, a number of oscillographic waveforms were photographed as shown in Fig. 9-68. Part (A) represents the voltage waveform from the plate of *V1* to ground. The sharp voltage pulse of this photo indicates that tube *V1* is cut off for only a short period of time. This means that tube *V2* is cut off for a longer period of time and has a plate current that flows in pulses of short time duration. The photo at (A), although being the voltage waveform at the plate of *V1*, can also be said to represent the picture of plate current for tube *V2*. This plate voltage measures 45 volts peak to peak.

The photo at (B) is the grid-voltage waveform of tube *V2*. The exponential sloping section of each cycle (that part with the slow rise time) represents the grid voltage during the charging of capacitor *C1* and the remainder of the cycle occurs during the discharge of the capacitor. This grid voltage measures 40 volts peak to peak. During the discharge of *C1*, the grid is driven positive and this positive swing is indicated by the pips on top of the waveform in part (B). The voltage waveform between cathode and ground, measured across the 500-ohm cathode resistor, is shown at (C). This voltage measures only 6 volts peak to peak, but is shown enlarged for the sake of clarity. The voltage waveform between the plate of *V2* and ground (i.e. across *C2*) is indicated at (D) and is seen to be the desired sawtooth signal. This output sweep signal measures 40 volts peak to peak.

The frequency of operation of the circuit under test was approximately 2,000 cps. To calculate the exact frequency of oscillations would involve too many circuit parameters. However, an approximate computation of the frequency can be made by using the following formula suggested by J. L. Potter:[8]

$$f = \frac{0.4}{(R1)(C1)} \qquad (9\text{-}25)$$

where *R1* and *C1* are as indicated in Fig. 9-67.

One-Shot or Triggered Multivibrators

The multivibrators discussed thus far are of the free-running type; that is, they do not require any external signal to set them into oscillations. Quite often, a circuit design requires the use of a multivibrator that can be initiated into oscillation at specific instants rather than having the circuit oscillate at all times. Such circuits do exist and are commonly known as ***one-shot, driven***, or ***triggered*** multivibrators. In brief, such a multivibrator requires the use of an external signal to produce the desired output signal. Without any external signal applied, the normal operation of the circuit is such that one tube of the multivibrator circuit has plate current flowing all the time and the other tube is always cut off. With the application of an external signal, which is usually in the form of a sharp pulse, one cycle of multivibrator oscillation occurs, and the circuit returns to its original static condition.

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[8] *loc. cit.*

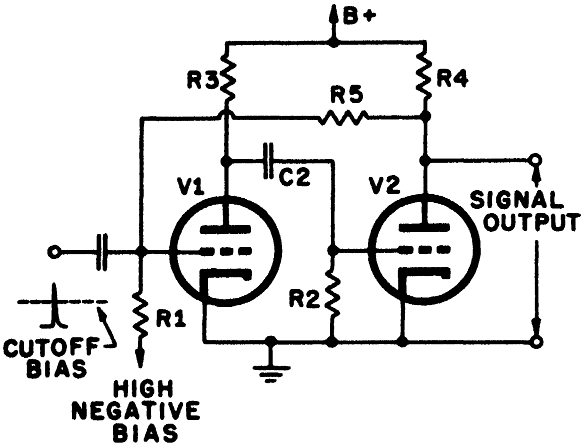

Fig. 9-69.—Typical schematic diagram of a one-shot (driven) plate-coupled multivibrator.

Let us first analyze a one-shot plate-coupled multivibrator, a typical circuit of which appears in Fig. 9-69. The primary difference between this circuit and that of the free-running circuit of Fig. 9-51 is that in the one-shot circuit, direct coupling, via resistor *R5*, is used from the plate of *V2* to the grid of *V1*, instead of capacitive coupling. Without any triggering pulse applied to *V1*, this tube is normally cut off and *V2* is normally conducting. In this state, a minimum positive potential from the plate of *V2* is applied to the grid of *V1*, and a high negative bias is applied to *V1*. The negative bias is much stronger than the small positive voltage, strong enough to cause *V1* to be permanently cut off. Remember that when *V2* has maximum conduction its plate potential is at a minimum, and when *V2* is not conducting (i.e. it is cut off) the plate potential is a maximum.

With a ***positive*** pulse applied to the grid of *V1*, which reduces the bias enough to drive the grid voltage above cutoff, tube *V1* will start conducting. Capacitor *C2*, which was previously charged from the B+ supply through *R3*, immediately discharges through the path consisting of *R2* and the low plate resistance of *V1*. The high discharging current of *C2* causes a high negative voltage drop across *R2*, biasing *V2* beyond cutoff; thus *V2* becomes nonconductive. The plate potential on *V2* in this nonconducting state is a maximum, and causes a high enough positive voltage to be applied to the grid of *V1*, via *R5*, to keep *V1* conducting.

The exponentially decreasing discharging current of *C2* will, after a certain amount of time, reach a value where the bias across *R2* will no longer cause tube *V2* to be cut off, and at this point *V2* starts conducting again. When this happens, the plate potential on *V2* decreases to its minimum value, and the fixed negative bias on the grid of *V1* is then strong enough to drive *V1* to cutoff. Thus the circuit is back again to its initial or normal state of operation. The single trigger pulse has brought about one cycle of oscillation of this multivibrator circuit. In order to initiate further oscillation of the circuit, the trigger pulse has to be applied again. The output signal is normally taken from the plate of *V2*.

Besides a positive pulse, it is also possible to have a negative pulse trigger a multivibrator. A driven plate-coupled multivibrator circuit using a negative triggering pulse is illustrated in Fig. 9-70. The primary difference between this circuit and that of Fig. 9-51 is that, in the one-shot circuit, the grid resistor of *V1* is returned to B+ instead of ground. Let us study this circuit to see how it operates. In the fixed state of operation, that is, without any trigger pulse applied, tube *V1* is normally conducting and tube *V2* nonconducting. The high negative bias on the grid of *V2* prevents it from conducting and the positive voltage on the

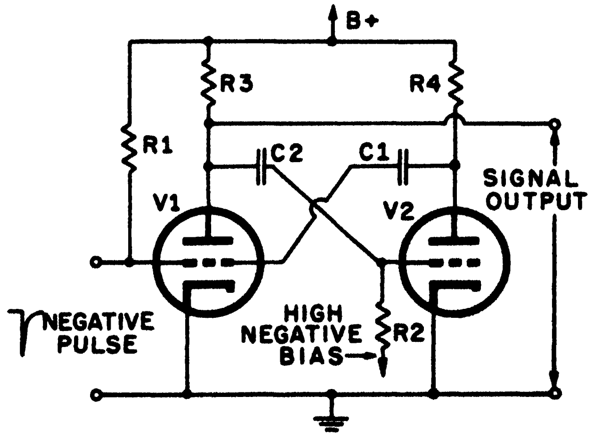

Fig. 9-70.—A driven plate-coupled multivibrator circuit employing a negative triggering pulse.

grid of *V1* keeps *V1* conducting. With a negative pulse applied to the grid of *V1,* the plate current flowing through the tube will decrease. This means that the potential on the plate of the tube will increase. This increased positive potential, which is in the form of a pulse, is applied to the grid of *V2* via *C2.* If the increase in potential at the plate of *V1* is strong enough so that the positive pulse from this plate will, for the moment, drive the grid of *V2* above its cutoff bias value, *V2* will start conducting.

Remember that the incoming negative trigger pulse need not cause *V1* to immediately cut off. It need only be strong enough to cause the resulting increased positive potential, which is in the form of a pulse, at the plate of *V1,* to overcome the bias on *V2* so that *V2* will become conducting. The moment this occurs, capacitor *C1,* which was originally charged from the B+ supply, starts to discharge through the low plate-resistance path of the now-conducting tube *V2,* and also through the grid resistor *R1.* The discharge circuit is completed through the power supply. The initial high discharge current of *C1* causes a negative voltage drop across *R1* which is strong enough to buck out the positive bias on *V1* and to drive this tube beyond cutoff. Thus we see that as soon as tube *V2* starts conducting, tube *V1* is cut off and we have the familiar switching action.

When *V1* stops conducting, the potential on its plate increases to a maximum; this maximum positive potential is applied to the grid of *V2* via *C2* and is strong enough to overcome the bias on *V2,* keeping it in the conducting state. As *C1* continues to discharge, its exponentially decreasing current will, at a point, produce a decreased negative voltage drop across *R1,* which, when combined with the positive voltage across *R1* from the plate supply, will cause *V1* to start conducting again. The moment that *V1* begins to conduct, the *V1* plate potential decreases, and the high bias on *V2* takes over again with *V2* becoming nonconductive. The circuit is now back to its original starting conditions with *V1* conducting and *V2* being cut off, and remains in this state. The initiation of the negative trigger pulse has caused the multivibrator circuit to produce one complete cycle of oscillation. In order to cause further oscillation, another negative pulse has to be applied to the grid of *V1.*

The circuits of Fig. 9-59 and 9-60 can be made to produce symmetrical or asymmetrical waveforms (i.e. balanced or unbalanced outputs), according to the value of component parts and applied voltages. In order to produce a sawtooth voltage, a capacitor is usually placed across the output of an unbalanced type of circuit, similar to the previously free-running multivibrator circuits. In this respect, remember that the tube across which the capacitor is placed should be conductive for only a small part of each cycle. This, of course, is to insure that the sawtooth waveform has a small retrace time.

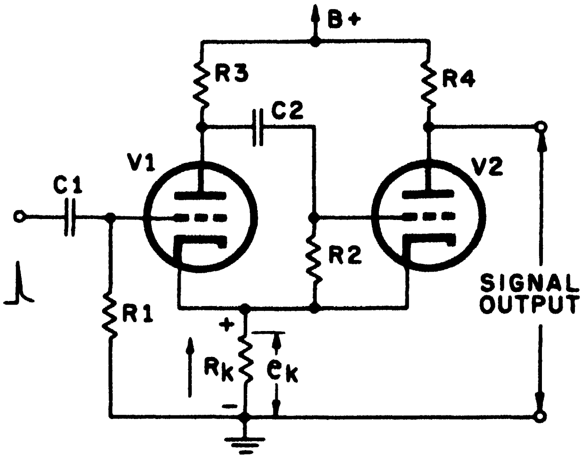

Fig. 9-71.—Schematic diagram of a driven cathode-coupled multivibrator.

So much for the triggered type of plate-coupled multivibrators. Let us now consider the cathode-coupled multivibrator when operated as a one-shot circuit, a typical circuit of which appears in Fig. 9-71. This particular circuit utilizes a positive pulse to initiate oscillation. Comparison of this circuit with that of Fig. 9-60 reveals that the chief difference is in the grid-return leads of the tubes. In Fig. 9-71, resistor *R2,* of the frequency determining *C2-R2* time-constant network, is returned to cathode, whereas in Fig. 9-60 the resistor, *R1* in that circuit, is returned to ground.

Without any external signal applied to the circuit of Fig. 9-71, but with the B supply and heaters voltages applied, tube *V1* is normally cut off and *V2* conducting. The grid of tube *V2* is normally at zero potential with respect to its cathode, because, generally speaking, no current flows through resistor *R2.* However, note that the grid of *V1* is at *ground* potential via resistor *R1.* The plate current of tube *V2* flows through the common cathode resistor R_k, and the voltage developed across this resistor puts a positive potential on the cathode of *V1* with respect to ground, as indicated in the drawing. This voltage drop, therefore, acts as the bias for tube *V1* and is usually high enough to cause tube *V1* to become cut off.

Refer to the waveform pictures of Fig. 9-72. Part (A) represents the pulse applied to the grid of *V1,* which is used to initiate the multivibrator circuit into oscillations. The waveform at (B) is the plate voltage

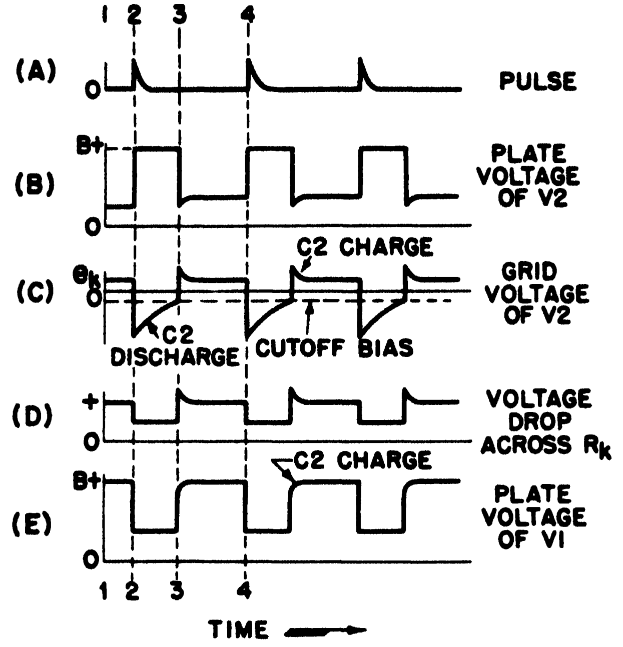

Fig. 9-72.—Waveforms produced under normal operating conditions by the driven cathode-coupled multivibrator circuit of Fig. 9-71.

of tube *V2,* that at (C) is the grid voltage of *V2,* at (D) is the cathode voltage across R_k, and the waveform of the plate voltage of *V1* is indicated at (E). In the following analysis of the circuit of Fig. 9-71, constant reference will be made to these waveforms.

From time *1* to the start of time *2,* no pulse is applied to the grid of *V1.* Thus between these two points, the plate voltage of *V2* is a minimum due to maximum plate current flow as indicated at (B) in Fig. 9-72. The grid voltage of *V2* with respect to ground is positive by the voltage drop e_k across the cathode resistor as indicated in the drawing at (C). The voltage drop e_k across cathode resistor R_k is a maximum due to maximum plate current of *V2* flowing through it. This is indicated at (D) of Fig. 9-72 during the time *1* to *2.* Since tube *V1* is not conducting, its plate voltage is at a maximum and equal to the *B*+ supply. This is indicated at part (E) of the waveform diagram.

Let us now apply a positive pulse to the grid of tube *V1.* The time this pulse is applied is indicated at *2* in Fig. 9-72. If this pulse is strong enough to drive the grid above its cutoff region, a number of actions occur which are considered to be simultaneous. First of all, tube *V1* starts conducting and the plate voltage decreases as shown at time *2* in part (E) of Fig. 9-72. As soon as *V1* starts conducting, capacitor *C2,* which was previously charged from the B supply, discharges through the low plate-resistance path of the now-conducting tube *V1,* and also through resistor *R2* to complete its path. The initial discharge-current flow is a maximum and is strong enough to cause a negative voltage drop across *R2* which biases tube *V2* beyond cutoff; see Fig. 9-72C. With tube *V2* not conducting, its plate voltage increases to a maximum, which is that of the B+ supply, as indicated in part (B) at time *2.* The voltage drop across R_k is a minimum, as shown at (D), because plate current of *V2* no longer flows.

All this occurs at time *2* in Fig. 9-72. At this time, the operation of the tubes is reversed, with *V1* now conducting and *V2* cut off. Tube *V1* continues to conduct as long as the negative voltage across *R2* keeps *V2* biased beyond cutoff. As *C2* continues to discharge, the discharging current decreases exponentially, which means that the voltage drop across *R2* decreases and hence the bias on *V2* also decreases. A point will be reached during the discharge of *C2* where the bias on *V2* will reach the cutoff value, and at this time *V2* will again start conducting. The action just described occurs during the time interval between *2* and *3* of Fig. 9-72. Once tube *V2* starts conducting, a number of so-called simultaneous actions occur. The current flowing through *V2* causes a voltage drop across R_k, which puts a negative bias on *V1,* reducing its plate current. This reduction of plate current causes the plate voltage of *V1* to increase; this increase is effectively coupled to the grid of *V2,* increasing the plate current of this tube. This means that the voltage drop across R_k will increase still further. This over-all action is cumulative, and a point will be reached where the voltage drop e_k exceeds the cutoff bias of *V1;* hence *V1* will no longer conduct. All these actions occur at time *3* in Fig. 9-72. As soon as *V1* stops conducting, capacitor *C2* again begins to charge up from the B supply. The effect of the charging of *C2* upon the waveform is noted in Fig. 9-72.

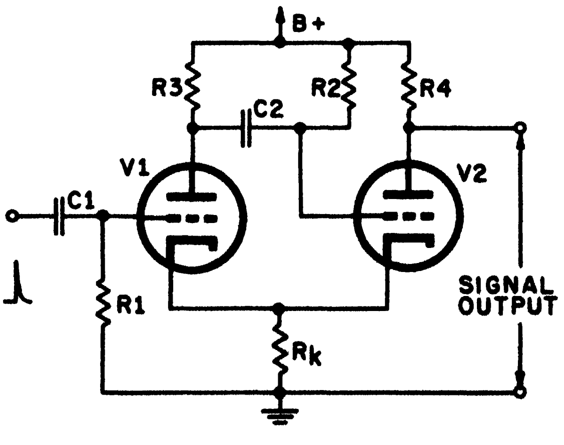

Fig. 9-73.—A slightly modified version of the one-shot cathode-coupled multivibrator circuit shown in Fig. 9-71.

If the trigger pulse is no longer applied to the circuit, then shortly after time *3* in Fig. 9-72, the circuit will return to its fixed operating condition. A trigger pulse is always needed to make the circuit oscillate. If continuous oscillation of the circuit is desired, the trigger pulse should be applied at periodic intervals. To indicate this, the pulse of curve (A) is shown in periodic form and the corresponding circuit waveforms are indicated below.

A slight variation of the one-shot cathode-coupled multivibrator of Fig. 9-71 is illustrated in Fig. 9-73. In this latter circuit, the grid resistor *R2* is returned to B+ instead of to the cathode of *V2*. In the normal condition, tube *V1* is nonconducting and *V2* conducting. Tube *V2* conducts because of the positive bias applied to its grid. Its plate current flows through cathode resistor R_k, thereby biasing both tubes. The choice of R_k and *R2* is such that the positive bias on the grid of *V2*, obtained from the B supply, is high enough to overcome the negative bias caused by the voltage drop across R_k, so that plate current continuously flows in *V2*. However, the voltage drop across R_k is such that it biases tube *V1* beyond cutoff.

When a positive pulse is applied to the grid of tube *V1*, having sufficient amplitude to overcome the cutoff bias of the tube, *V1* will start conducting and its plate voltage will begin to decrease. Since the voltage which existed across capacitor *C2* cannot change as rapidly as the change in plate voltage, this plate-voltage decrease causes a proportional drop in the grid-to-ground voltage of *V2*. This, in turn, causes a decrease in the plate current of *V2*, and, therefore, a decrease in the voltage drop across R_k. This further reduces the bias on *V1*, causing its plate current to increase still more and its plate voltage to decrease. Hence the grid-to-ground voltage of *V2* will decrease still further. Although the voltage drop across R_k is being reduced, due to the decrease in the plate current of *V2*, remember that the increasing plate current of *V1* also flows through R_k. However, the decrease in the plate current of *V2* is greater than the increasing plate current of *V1*. Capacitor *C2*, which was charged from the B supply, starts to discharge through the low plate resistance of the now-conducting tube *V1* and also through R_k and *R2*. All these actions, up to the initial discharge of *C2*, occur simultaneously. The discharge current of *C2* causes a voltage drop across *R2* which makes the grid of *V2* more negative than its cutoff value, and *V2* no longer conducts.

As *C2* continues to discharge, the decreasing discharge current causes the voltage across *R2* to likewise decrease. A point is reached where the grid voltage of *V2* with respect to cathode is above the cutoff value, and *V2* begins to conduct again. When this happens, *V1* again becomes nonconductive because of the following action. The voltage drop across R_k increases because of the plate current of *V2* increasing, the plate current of *V1* then decreases, the effective voltage at the grid of *V2* increases, the current through *V2* increases further, and the drop across R_k increases still more. This action continues until *V1* is cut off because of the high bias due to the voltage drop across R_k, and the circuit returns to its normal state. The waveforms of operation are similar to those of Fig. 9-72.

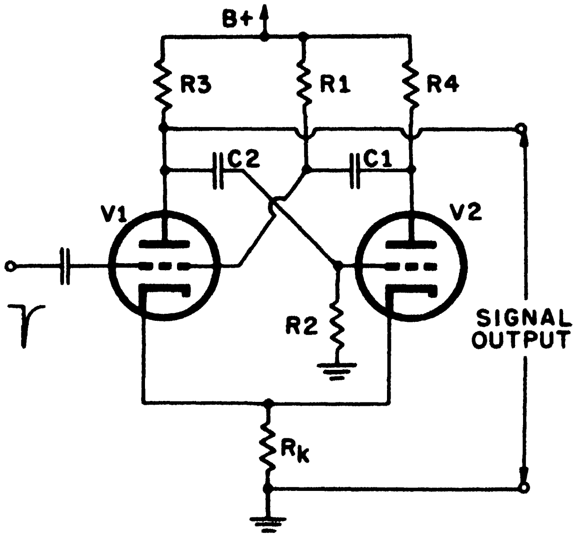

Fig. 9-74.—Circuit of a cathode-coupled multivibrator using a negative triggering pulse.

A cathode-coupled multivibrator circuit which uses a negative pulse to trigger the circuit is shown in Fig. 9-74. In this circuit, two coupling capacitors and grid resistors are used, with resistor *R1* being returned to B+ and *R2* to ground. The normal operating conditions of the circuit are such that the positive voltage on *V1* via *R1* keeps it conducting. The plate current of *V1* flowing through R_k is great enough to cause a voltage drop across R_k such that the bias on *V2* is beyond its cutoff value and, therefore, *V2* is nonconductive. Let us examine the operation of this circuit when a negative pulse is applied to the grid of tube *V1*.

This pulse will cause the plate current of *V1* to decrease and its plate voltage to increase. This change in plate voltage, which is in the form of a pulse, appears as an increased voltage at the grid of *V2*. The input trigger pulse must be strong enough to make the positive pulse at the plate of *V1* of sufficient magnitude to overcome the effective bias on *V2* and cause the tube to conduct. With *V2* conducting, its plate voltage is reduced. This reduced voltage is coupled to the grid of *V1* via *C1* and causes a reduction in the voltage between the grid and ground of *V1* which further reduces the current through this tube. These over-all actions continue until tube *V1* is cut off. While this is happening, the voltage across R_k also decreases. This is so because the increase in plate current of *V2* is smaller than the decrease in plate current of *V1*.

As soon as tube *V2* starts conducting, capacitor *C1* starts discharging through the low plate resistance of tube *V2* and also through *R1* and *R4*. The initial discharge current is a maximum and the voltage drop it causes across *R1* is more than enough to keep *V1* biased beyond cutoff. As *C1* continues to discharge, the bias on *V1* will gradually be reduced; when it reaches the cutoff value of the tube, *V1* will start conducting again. When this happens, cumulative variation in the plate voltages and currents of each tube occurs, similar to that previously discussed, with the result that tube *V2* once again becomes cut off due to the increase in voltage across R_k. Thus we are back to the original starting conditions. In order to cause oscillation once again, another negative pulse has to be applied to the grid of *V1*.

Comparison between the one-shot cathode-coupled multivibrator of Figs. 9-71 (with the grid of *V2* returned to cathode) with those of Figs. 9-73 and 9-74 (with the grids returned to B+) brings to light an interesting feature. In Fig. 9-71, the time-constant network which determines the frequency of operation is *R2-C2*. The discharge of this capacitor is represented by the exponential curve shown in part (C) of Fig. 9-72. An enlarged portion of this discharge curve is redrawn in Fig. 9-75A. If capacitor *C2* were to become fully discharged, its representative curve would approach the cathode potential, as indicated by the dashed part of the exponential curve. When the cutoff bias of the tube is reached, however, the tube starts conducting and the capacitor stops discharging. According to this curve, the capacitor has discharged a major portion of its original charge which results in the grid-voltage waveform between points *1* and *2* having great curvature. This means that the cutoff voltage of the tube of Fig. 9-71 is approached very gradually because the amplitude of the upper portion of an exponentially rising curve changes very little with respect to time. Such a type of multivibrator circuit will *not* insure very accurate timing and frequency stability, because variations in tube performance will cause appreciable changes in the time at which the discharge curve will reach the cutoff voltage of the tube.

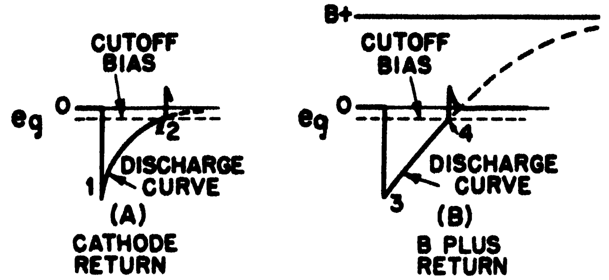

Fig. 9-75.—An enlarged portion of the capacitor discharge curve shown in part (C) of Fig. 9-72 for the cathode-return type driven multivibrator is given in (A). Compare this with the grid-voltage curve of the positive-grid-return type multivibrator shown in (B).

With the positive-grid return circuits of Figs. 9-73 and 9-74 this disadvantage does not manifest itself. In Fig. 9-73, the chief frequency-determining time constant is *R2-C2* and for Fig. 9-74 it is *R1-C1*. In each circuit, the grid resistor of the time-constant networks under consideration is returned to the B supply. Consequently, as the capacitor discharges, which is usually represented by the grid-voltage waveform, it would approach the B supply value if it became fully discharged. This is best illustrated by the grid-voltage curve of Fig. 9-75B.

As the capacitor discharges, a point will be reached, point *4* in the drawing, where the grid voltage will be at the cutoff value of the tube and conduction will occur, thereby preventing further discharge of the capacitor. This drawing indicates that only the very beginning of the discharge curve, between points *3* and *4*, is effective. This, as you can see, is the most linear part of the discharge curve. It means that the cutoff value of the tube is approached very rapidly due to the sharp slope of the curve at this point as compared to that of curve (A). Therefore, changes in tube performance will not cause appreciable changes in the time that the discharge curve will reach the cutoff value of the tubes. Consequently, we see that for more stable operation, it is best to employ the cathode-coupled multivibrator with positive-grid return rather than the one with the grid returned to the cathode.

In all of the one-shot or triggered cathode-coupled multivibrator circuits, the use of a capacitor across the output will produce a sawtooth waveform output. For the sawtooth to have a rapid flyback time, the conduction time of the tube across which the capacitor is placed should be very small compared to its period of nonconduction.

One-shot or triggered multivibrators are also used in oscilloscopes as gate circuits in supplying pulses for the triggered sweep system of the oscilloscope. For example, in Du Mont Type 248 oscilloscope (see

Chapter 22), a sync pulse is used to activate a one-shot cathode-coupled multivibrator. A special switching system permits choice of the rate of pulse output from this circuit. The pulse output from this cathode-coupled multivibrator is used to drive the triggered sweep circuit of the oscilloscope.

Blocking Oscillators

The third common type of circuit used to produce a linear sawtooth waveform is the *blocking oscillator.* The blocking oscillator is also a relaxation oscillator of *R-C* variety that employs a hard tube. Similar to the multivibrators, blocking oscillators can be either free-running or of the triggered type. In general, such circuits are considered as pulse generators because of the waveshapes available from them. There are a few different types of blocking-oscillator circuits, but the purpose of each is usually the same, to produce a pulse output. Many pulse-forming circuits can easily be put to use to help produce a sawtooth output voltage. Let us study a common type of blocking-oscillator circuit to see how it operates. A typical free-running circuit appears in Fig. 9-76.

The transformer employed in such a circuit is one quick means of distinguishing the blocking-oscillator type of relaxation oscillator from others. This circuit is very similar to the usual type of tickler feedback oscillator commonly employed in radio receivers. In this latter type of circuit, the oscillator is usually tuned by a variable capacitance across the grid coil. In the blocking oscillator, the distributed capacitance of the coil, interelectrode capacitances of the tube, and stray wiring capacitances take the place of the variable capacitance of the tuned-grid tickler feedback oscillator. Therefore, the blocking oscillator can be considered a form of tickler-coil oscillator with a high *L* to *C* ratio in its grid circuit.

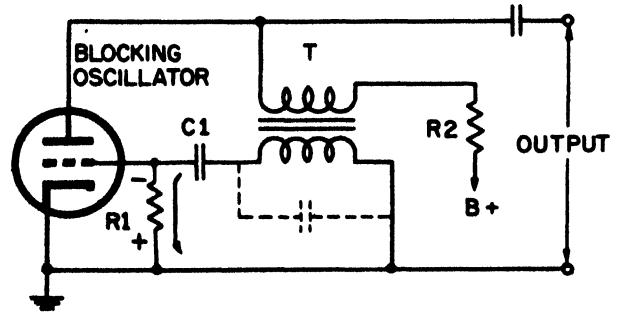

Fig. 9-76.—Typical circuit of a free-running blocking oscillator.

In the tickler oscillator, the time constant of the grid-coupling capacitor and resistor (*C1-R1* if Fig. 9-76 is considered as a tickler oscillator) is so chosen that, in conjunction with the tuned circuit, the tube is operated class C. With the *C1-R1* combination providing grid-leak bias, the oscillator generates a continuous sine wave at the resonant frequency of the circuit. In the blocking-oscillator circuit, the time constant of *R1-C1* is deliberately chosen to be very large so as to interfere with or *block* the continuous production of sine-wave oscillations. Thus we see the reason why the circuit is called a blocking oscillator.

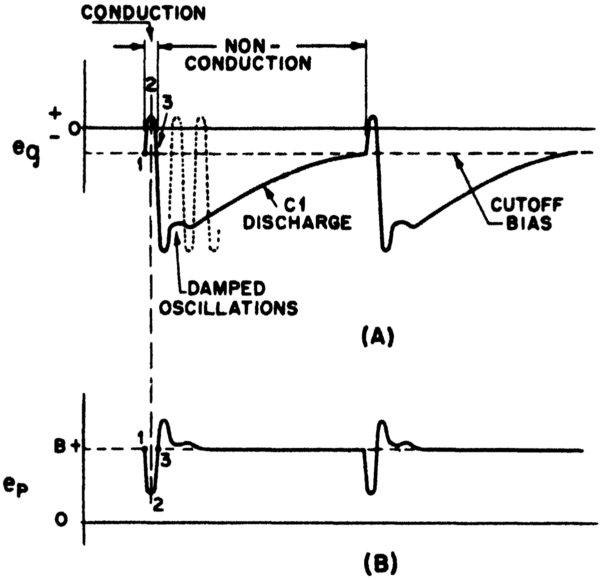

Fig. 9-77.—Operating waveforms produced by the blocking-oscillator circuit of Fig. 9-76.

The inductive coupling between the plate and grid circuit of the blocking oscillator is very tight and the transformer turns ratio is such that the signal coupled to the grid will be very strong. The large time constant of the *R1-C1* combination is much greater than the time required to complete one cycle of operation of the tuned-coil arrangement. Because of these factors, a high bias is developed and prevents the circuit from oscillating for any appreciable period of time. Typical waveforms of operation of the blocking oscillator are shown in Fig. 9-77. The operation of the blocking oscillator of Fig. 9-76 will be more readily understood by continuous reference to these waveforms.

Let us consider the circuit just as the supply voltage is turned on after the tubes have been warmed up. It should be remembered that this circuit is regenerative, the transformer being connected so that any change of current in the plate circuit will produce a change in the grid circuit which will aid the original change. The application of the supply voltage will cause current to flow in the plate circuit of the tube. Since no current was flowing previously, this current now flowing is considered as a current change. This change in current flows through the plate winding of the transformer and is coupled to the grid of the tube via the grid transformer winding. Since the circuit is regenerative, the voltage that is coupled to the grid must aid the original increase in plate current. Therefore, the polarity of this grid voltage is such as to make the grid positive with respect to ground and hence also to cathode. This increase in grid potential causes an increase in the plate current; this plate-current change is again coupled to the grid as a higher positive voltage, still further increasing the plate current. This increase in grid voltage is shown at the beginning of waveform (A) of Fig. 9-77. The increasing of plate current continues. As in the regular tickler oscillator, this positive grid voltage causes grid current to flow through *R1,* with the result that electrons begin to pile up on the grid side of capacitor *C1,* thereby charging it. The charging path of the capacitor is through the transformer secondary and the low grid-to-cathode resistance of the tube. This charging process continues as long as grid current flows.

The circuit impedances and tube characteristics are such that a saturation point will be reached where the plate current and hence grid voltage will increase no more. This is indicated at point *2* in Fig. 9-77. When this maximum point has been reached, the plate current starts to decrease. This means that a change in plate current once again occurs, but in a decreasing direction, and, since the circuit is regenerative, a decreasing voltage is transformer-coupled to the grid. This reduces the positive potential on the grid and also reduces the plate current. This further change (i.e. reduction in plate current) causes the grid voltage to likewise reduce still more. This over-all action continues until the grid is driven below its cutoff value.

Once the grid voltage is no longer positive grid current stops flowing and the electrons which have piled up on the capacitor look for a ready path to leak off. The capacitor discharges through the high-valued grid resistor *R1* because the low-resistance path from cathode to grid no longer exists, the grid-current flow having ceased. Before we examine what happens during the discharge of *C1* through *R1*, let us study the formation of the waveforms up to this point.

First of all, it is seen that as the plate current is increasing, the plate voltage is decreasing to point *2* in Fig. 9-77B. We also know that the grid voltage rises to a maximum positive value as the plate voltage decreases to a minimum value. As the grid voltage starts decreasing and reaches the cutoff value of the tube, the plate voltage increases to the B+ value, since no plate current flows when the tube becomes cut off. The time that it takes the grid voltage at (A) or the plate voltage at (B) to go from points *1* to *3* is primarily determined by the effective inductance of the grid coil, its distributed capacitance, the tube interelectrode capacitance, and the coil resistance. In short, it is often stated that the periods of these pulses represent one half-cycle of oscillation of the resonant frequency of the grid circuit.

Let us now refer back to the time when *C1* starts discharging. In the ordinary tickler oscillator, the choice of *R1* is relatively small and the time constant of *R1-C1* of a small enough magnitude so that the charge on *C1* can leak off fast enough to have the circuit continue to oscillate. In other words, the initial pulse between points *1* and *3* in Fig. 9-77 represents one half-cycle of oscillation in the normal tickler-coil oscillator circuit, with the frequency of oscillation being determined by the so-called tuned circuit of the system. The continuation of the oscillations is indicated in Fig. 9-77A by the sine waves drawn in dotted form.

In the blocking-oscillator circuit, tht grid resistor *R1* is made very high so that, in conjunction with *C1,* a long time constant is represented. This long time constant means that the capacitor will take quite some time to discharge through the high-valued resistor. In actual practice, the product of *R1-C1* is made much larger than the period represented by the half-cycle of oscillation of the effective tuned-grid circuit. The *discharge* current flows through *R1* in the direction indicated in Fig. 9-76 and causes a negative voltage to be applied to the grid of the tube. This negative voltage is a maximum at the initial moment of discharge.

The effective tuned circuit of the transformer is immediately set into oscillations by the first surge of plate current flowing through the transformer windings. The initial half-cycle of oscillation and the rise and decay of plate current occur very rapidly. When the grid voltage is reduced beyond its cutoff value, which is at the end of the first half-cycle of transformer oscillation, the tuned circuit which is already energized continues to oscillate. However, primarily due to the physical construction of the transformer, which normally has a low *Q,* and the fact that the tube becomes cut off and does not supply any more energy to the transformer, these oscillations are quickly damped out. This is indicated in the waveform drawings of Fig. 9-77. In some circuits, a resistor is shunted across the grid winding of the transformer to decrease the *Q* and further dampen oscillations. In most practical circuits, the oscillations are immediately damped out so that after the grid is driven beyond cutoff, as capacitor *C1* discharges through *R1,* the tuned circuit no longer oscillates.

As *C1* continues to discharge, the negative voltage on the grid decreases and does so in exponential fashion as seen at (A) in Fig. 9-77. Since *R1* is a much higher resistance than the grid-to-cathode resistance of the tube when the grid is positive and conducting, a much longer time is required to discharge *C1* than to charge it; hence the grid remains negative for a much longer period of time than it was positive. A time will be reached during the discharge of *C1* when the negative voltage on the grid will be reduced to just above cutoff and plate current will start flowing once again. At the moment this happens, the original operation starts over again where the tuned circuit is forced into oscillations, the grid voltage starts increasing in the positive direction, and *C1* again begins to charge when the grid goes positive.

The curve at (B) in Fig. 9-77 is representative of the plate-voltage waveform and also the signal output. The output voltage is in the form of periodic pulses which are seen to be above and below the *B+* reference axis. The formation of the pulses below the *B+* axis is due to the increasing and decreasing plate-current flow as explained previously. The pulses above the *B+* line are formed when the tube is already cut off and plate current no longer flows. These pulses are primarily due to the energy stored in the plate winding of the transformer and are considered as the beginning of the damped oscillations of the tuned circuit.

The frequency of the pulse repetition of the waveforms of Fig. 9-77 is primarily determined by the *R1-C1* combination for a given system. Other factors, such as the tube used, characteristics of the transformer employed, and circuit voltages, also determine the frequency of operation of the free-running blocking oscillator, but the values of *R1* and *C1* are usually the variable ones.

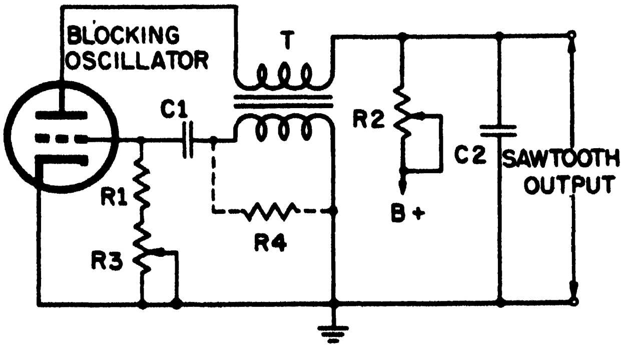

Fig. 9-78.—A simple type of blocking-oscillator circuit for producing a sawtooth signal output.

After understanding the basic operation of the blocking oscillator, it is a simple matter to figure out a method of obtaining a sawtooth output. In the asymmetrical multivibrator circuit, a capacitor is placed across the output circuit of the tube that has the smaller period of conduction. In the blocking oscillator, the tube by the nature of its circuit arrangement conducts for a much shorter period of time than when it is cut off. Placing a capacitor across its plate circuit produces the same result. Appearing in Fig. 9-78 is a simple blocking oscillator circuit used to produce a sawtooth signal output. It is the addition of capacitor *C2* which helps produce the sawtooth wave. In brief, the sawtooth signal is produced as follows:

During the nonconduction period of the tube, *C2* charges up through *R2* from B supply. The charging is exponential, but the choice of value of *C2* and *R2* is such that only the beginning of the exponential charge curve is utilized, in order that the sawtooth output signal be as linear as possible. As soon as the tube starts to conduct, it represents a very low resistance across *C2* and the capacitor discharges through this path. Since the period of conduction is very short, the flyback time of the sawtooth output signal, which is represented by the discharge of *C2,* is very rapid. Resistor *R3* is inserted to indicate how the frequency of such a circuit may be controlled. Blocking oscillators are seldom used as the sweep generator in oscilloscopes, but are often used in television receivers. When blocking oscillators are used in the horizontal or vertical oscillator sweep circuit in television receivers, resistor *R3* is usually termed the "hold" control (actually "horizontal hold" or "vertical hold," as the case may be). Resistor *R4* is inserted to illustrate its use as a "damper" to the oscillations produced by the effective tuned transformer circuit. Resistor *R2* is made variable so that it can serve as an amplitude control.

Blocking Oscillator and Discharge Tube

A disadvantage exists in the blocking-oscillator circuit of Fig. 9-78 in reproducing the sawtooth signal. From this drawing, you will note that the discharge circuit of *C2* includes the plate resistance of the tube in series with the plate winding of transformer *T*. Therefore, when capacitor *C2* discharges, the discharge current must flow through the winding. The inductance of this coil has a pronounced retarding effect upon the rate of discharge of *C2* by acting as an additional impedance in the discharge path. Therefore, if the discharge is to take place rapidly, only a partial discharge of *C2* will be effected in the short period that the tube

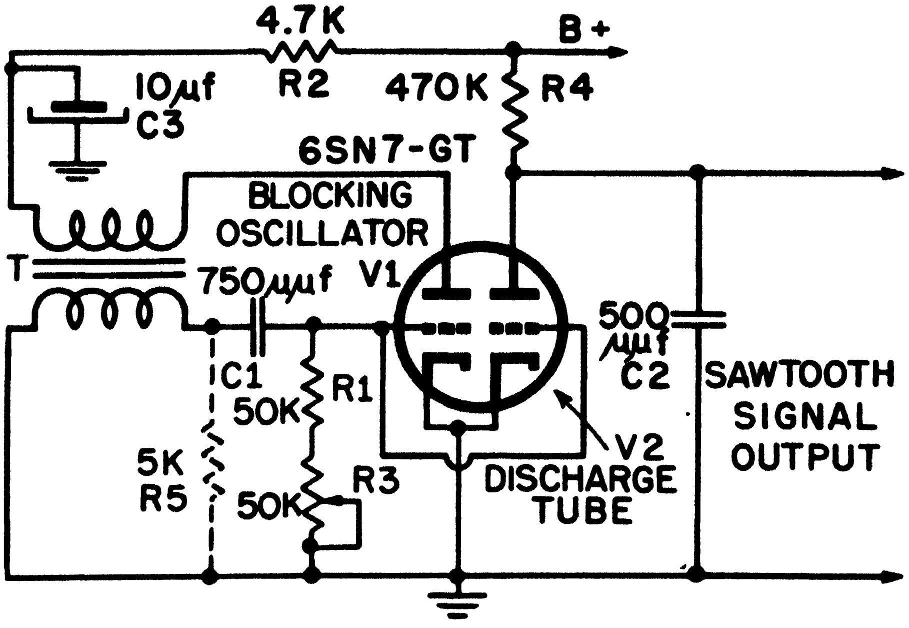

Fig. 9-79.—Schematic diagram of a blocking oscillator used in the sweep circuits of television receivers employing a discharge tube.

conducts. This results in a sawtooth wave of very low amplitude. Of course, the amplitude of the wave can be controlled somewhat by variation of *R2*, but the retarding effect due to the plate coil nevertheless exists. On the other hand, if a large amplitude is desired in this circuit, the retrace time of the sawtooth will be too great because the charge and discharge of *C2* must cover a large voltage range. To circumvent these undesired features, a second tube may be used as a circuit "switch" to charge and discharge the capacitor across which the sawtooth waveform is produced. Such a tube is often called a *discharge tube.*

When blocking oscillators are employed as the sweep circuit of television receivers, they invariably use the discharge tube. A typical circuit is illustrated in Fig. 9-79. Two separate tubes may be used or a duo-triode tube such as that illustrated. Tube *V1* is the blocking oscillator and *V2* is the discharge tube. The cathodes of both tubes are grounded and the grids of each are connected together. Because of this latter connection, whatever signal is present at the grid of the blocking oscillator is also present and in phase at the grid of the discharge tube. The blocking oscillator circuit is essentially the same as before. Capacitor *C1* and resistors *R1* and *R3* are the chief factors in controlling the repetition frequency of the pulses and hence the frequency of the sawtooth output signal. The plate of *V2* is connected to *B+* through *R4*, which, in conjunction with *C3*, forms a decoupling network.

The circuit is so arranged that when the grid of *V2* is driven beyond cutoff by the negative pulse on the grid *V1*, plate current no longer flows in *V2*. Capacitor *C2*, therefore, begins to charge up through resistor *R4*. As soon as the discharging current of *C1* causes the negative voltage on the grid of *V2* to reach cutoff, the tube will begin conducting. (Usually the circuit is so designed that the cutoff voltages of both grids are the same, so that both tubes become conducting and nonconducting at the same time.) Once *V2* conducts, *C2* begins discharging through this tube because its plate resistance is a very low impedance path across *C2*. You will note that during the discharge of *C2*, the discharge path is through tube *V2* only, with no inductance included; therefore, most of the charge on *C2* is removed during the conduction period of *V2*. The result is a higher amplitude sawtooth wave with a rapid retrace time.

Let us examine some of the waveforms of operation from the circuit of Fig. 9-79. This circuit is designed to operate at the horizontal sweep frequency of television receivers. This frequency is actually 15,750 cps, but by varying potentiometer *R3*, the repetition frequency can be changed. This control, as mentioned previously, is termed the "horizontal hold". A number of photographs illustrating the operation of this circuit appear in Fig. 9-80. An oscillogram of the signal at the grid of either triode is shown in part (A). Compare this picture with the drawing of Fig. 9-77 and notice how similar they are. In the photograph, more oscillations of the transformer have occurred, as seen by the many ripples along its discharge curve. The signal at the plate of tube *V2* is illustrated by the photo at part (B). Note the oscillations due to resonance in the transformer circuit. The similarity between this photo

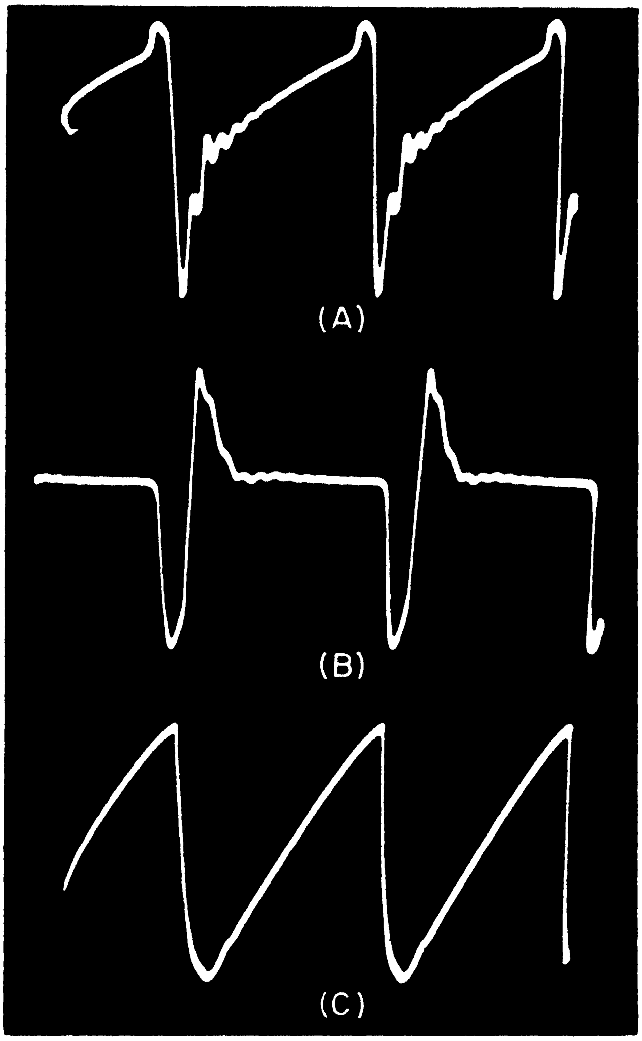

Fig. 9-80.—Oscillograms of waveshapes produced by the blocking-oscillator circuit of Fig. 9-79; the signal at the grid of either triode is shown in (A), the signal at the plate of tube *V2* in (B), and the output sawtooth waveform of the discharge tube is shown in (C).

and that of the drawing of Fig. 9-77B is readily apparent. The oscillogram for the output sawtooth waveform taken at the plate of the discharge tube is shown at part (C). You will note that the bottom of the rising portion of the wave is not very linear. This nonlinearity is due to the transformer oscillation, especially the first "overshoot" due to these oscillations.

The transformer oscillations which appear in the oscillograms of Fig. 9-80 occur for a longer period of time than is desired, and result because the Q of the blocking oscillator transformer is not low enough. The effect of a low Q on the damping out of these undesired oscillations can easily be demonstrated by inserting a damping resistor across the transformer secondary. This resistor is designated as *R5* and shown in dashed form in the circuit of Fig. 9-79. The resulting oscillograms of the grid and plate voltage are shown at Fig. 9-81A and B, respectively. The use of damping resistor *R5* has reduced the amplitude of the resulting waveforms and also the frequency of operation. These changes, which can be considered as by-product effects, are not critical inasmuch as each can be readjusted. In the grid-voltage waveform at (A), the majority of the transformer oscillations are seen to be damped out. Compare this photo with that at Fig. 9-80A and notice the difference. The plate-voltage waveform at Fig. 9-81B has the oscillations completely damped out. Notice the difference between this photo and that of Fig. 9-80B.

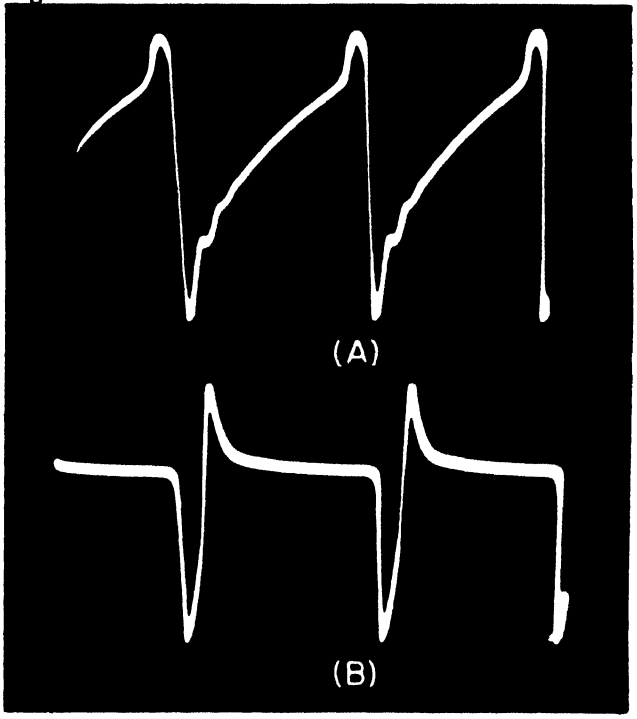

Fig. 9-81.—Oscillograms of the grid and plate voltages of the blocking oscillator of Fig. 9-79 after insertion of a damping resistor (*R5* in Fig. 9-79) across the transformer secondary, are shown in (A) and (B), respectively.

Other Blocking-Oscillator Circuits

There are a number of variations in the blocking-oscillator circuits thus far illustrated. One such variation employs a three-winding transformer, the circuit of which appears in Fig. 9-82. If for the moment we were to neglect coil *L3*, the operation of this circuit would be exactly the same as that of Fig. 9-76, assuming, of course, that both circuits are designed to operate as blocking oscillators. In Fig. 9-76, the output pulse is taken across the plate circuit, and from the waveform of Fig. 9-77B it is seen that the main pulses are negative in character. By coupling another coil, *L3* in Fig. 9-82, to the plate coil, the output from the plate circuit can also be made to appear across this new coil. Coil *L3* is wound so that the pulse appearing across *L3* is

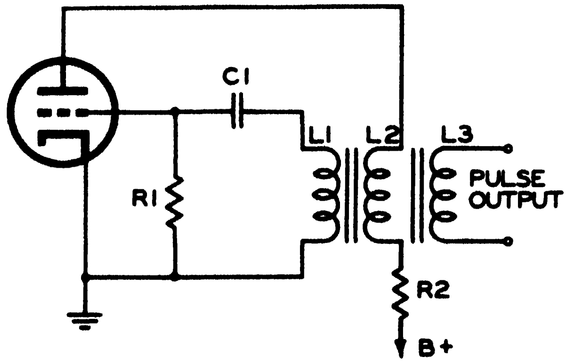

Fig. 9-82.—Modified version of a blocking-oscillator circuit employing a three-winding transformer.

180° out of phase with that across *L2*. With this type of circuit, there will be a positive pulse output. Placing a capacitor across *L3* will not produce a sawtooth signal output as in the other systems.

Like multivibrator circuits, blocking oscillators can also be used as driven pulse generators. A typical triggered blocking oscillator is shown in Fig. 9-83A. The grid of the tube is returned to a source of negative bias so that, without any trigger pulse applied, the tube is normally cut off and there is no plate-current flow. Capacitor *C1* is charged by the negative bias voltage through *R1*. Let us see what happens when a strong positive trigger pulse is applied to the grid circuit.

If the pulse is strong enough to overcome the high cutoff bias, the tube will start conducting. Once this happens, the circuit begins to function similarly to the free-running blocking oscillator. The regenerative effect of the transformer coupling between the plate and grid increases the plate current to a maximum, the plate voltage to a minimum, and the grid voltage to a maximum positive value. After these peak values have been reached, they reverse in direction. With a decrease in plate current, the grid voltage decreases to a negative value beyond the fixed bias on the tube. During the time that the grid becomes positive, grid current flows and capacitor *C1* is charged still further. In other words, before the negative charge on *C1,* due to the bias, has a chance to leak off, the grid current (which flows during a very short interval) causes the capacitor to become charged to an even greater extent in the same negative direction. As soon as the grid voltage has decreased to its maximum negative value, the capacitor begins to discharge through *R1,* the transformer secondary, the trigger-pulse source, and the bias supply. The impedance of the trigger source is assumed to have negligible effect on the discharge time constant.

The waveshape during discharge is similar to that of Fig. 9-77A. At the end of the *C1* discharge, the grid voltage has increased from its maximum negative value to that of the applied bias on the tube. Where this point is reached, one cycle of operation is complete and the circuit is restored to its normal nonconducting condition. Another positive trigger pulse is required to cause oscillation again. The signal output from the driven circuit of Fig. 9-83A is similar in shape to that of Fig. 9-77B. A sawtooth voltage output can be obtained by use of a capacitor across the output circuit.

The circuit of Fig. 9-83B is a driven blocking oscillator which uses a discharge tube to produce a sawtooth-wave output. When the grid of *V1* is driven above its cutoff bias value, both tubes conduct; capacitor *C2,* which had previously charged from the B+ supply through *R4,* then discharges through the low plate-resistance path of conducting tube *V2.* When both tubes are nonconductive, *C2* charges again.

A negative pulse can also be used to trigger the circuits of Fig. 9-83. This negative pulse would be applied to the plate circuit of the oscillator. Due to the transformer coupling, a positive pulse would appear at the grid of the tube, causing conduction by driving the grid above its cutoff value.

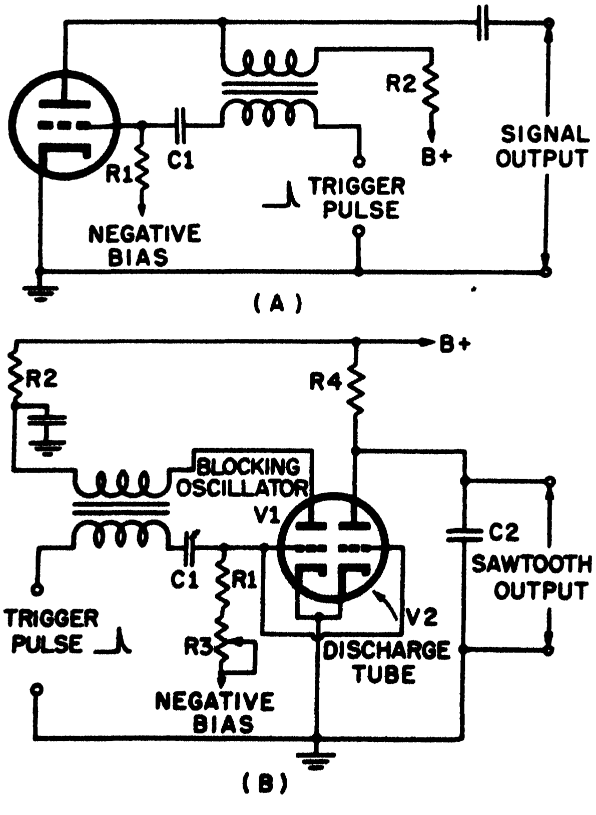

Fig. 9-83.—Typical circuit of a triggered blocking oscillator.

Pulse Shape and Other Considerations

In the analysis of blocking oscillators thus far, we have not said very much about the exact shape of the pulse output, except to mention something about sinusoidal oscillations. In fact, we went so far as to illustrate the pulses as being the beginning of sine waves (see Fig. 9-77). In reality, the action of the blocking oscillator, because of the use of an iron-core transformer, is such that the pulses are not sinusoidal. Close scrutiny of the oscillograms of Figs. 9-80 and 9-81 will indicate the nonsinusoidal shape of these pulses. In order to have a better visualization of these pulses, the waveforms of Fig. 9-80A and B have been spread out, and one cycle of each is shown in Fig. 9-84A and B, respectively. From these latter photos, it will be seen that the pulses are not sinusoidal. The explanation of the pulse shaping is best understood by a physical analysis of the transformer action involving magnetomotive force and magnetic flux.[9] However, the exact shape of the pulse is not very important to us, since it has very little effect on the shape of the resulting sawtooth signal.

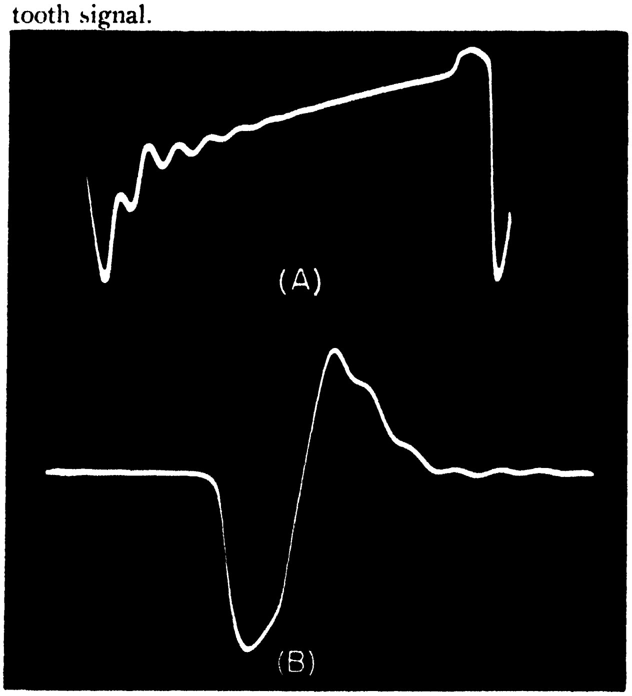

Fig. 9-84.—One cycle each of the waveforms shown in Fig. 9-80A and B are repeated here in expanded form in (A) and (B), respectively.

Let us refer to the grid-voltage waveform of Fig. 9-77A. Notice that if the small amount of oscillation caused by transformer resonance did not exist, this grid voltage would then roughly represent a sawtooth wave. If the circuit were so designed that the actual discharge of the grid capacitor only occupied a small part of the complete discharge curve, the resulting grid waveform would have a fairly linear characteristic. One method of accomplishing this is to return the grid resistor to $B+$ instead of to ground. This is what is actually done in some blocking-oscillator circuits. Such blocking oscillators are employed in sawtooth sweep circuits of British television receivers; the simplified diagram of a typical circuit appears in Fig. 9-85. Note in this circuit that the *R1-C1* network appears on the side of the transformer secondary away from the grid (compare with Fig. 9-78), with *R1* connected to $B+$ and the output sawtooth signal taken across *C1*.

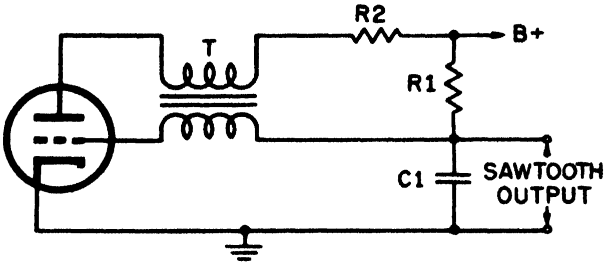

Fig. 9-85.—Simplified circuit diagram of a modified blocking oscillator, where the grid resistor is returned to $B+$ instead of the conventional return to ground.

Utilization of such a circuit as a sawtooth generator is not practicable unless oscillations due to transformer resonance with the lumped circuit capacitances are almost completely eliminated. One method of accomplishing this, besides damping, is to neutralize the capacitance with which the transformer oscillates.[10] However, complete neutralization is not usually possible, because the transformer always possesses some leakage inductance.

Blocking oscillators are rarely used in oscilloscopes as the linear sweep circuit because the transformer oscillations and overshoots are very prominent at the higher frequencies, and the resultant sawtooth is, therefore, not very linear. At low and medium frequencies, the sawtooth output of blocking oscillators is more useful. The neutralization method mentioned above (applicable to the type of circuit shown in Fig. 9-85 or to the other circuits previously discussed) is one method whereby blocking oscillators can be made more useful at higher frequencies.

At the beginning of this chapter, we discussed a very simple method for checking the linearity of the sawtooth sweep in oscilloscopes. A square-wave signal was first differentiated and the resulting pulses were then fed to the vertical amplifier of the oscilloscope or

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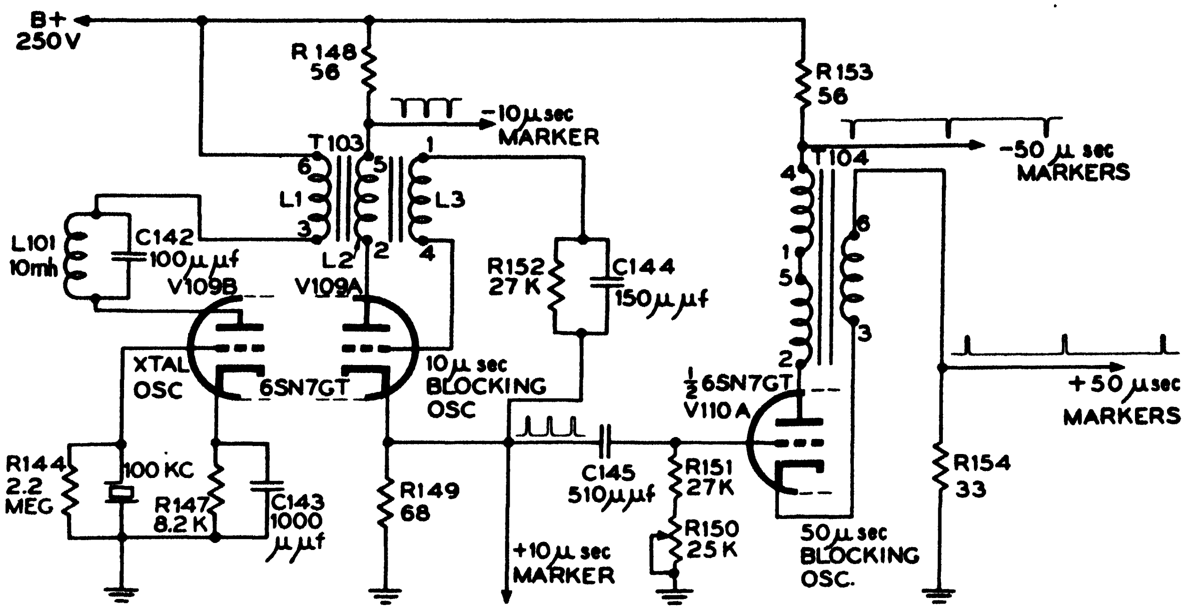
[9] Cruft Electronics Staff, "Electronic Circuits and Tubes," 1st ed., pp. 804-809, McGraw-Hill Book Co., Inc., New York, N. Y., 1947.

[10] Cocking, W. T., "Blocking oscillators," *Wireless World*, vol. 55, pp. 230-233, June, 1949.



After Du Mont Labs.

Fig. 9-86.—Schematic diagram of the marker generator used in the Du Mont Type 256D oscilloscope; here the pulse output from the crystal oscillator is used to trigger the blocking oscillator.

to grid of the cathode-ray tube. Blocking oscillators are often employed in oscilloscopes for checking linearity and also as marker generators. For example in the Du Mont Type 256D oscilloscope, the pulse output from a crystal oscillator is used to trigger a blocking oscillator similar to the type indicated in Fig. 9-82. The schematic for this circuit is illustrated in Fig. 9-86. Tube *V109* acts as the crystal oscillator with the crystal frequency being 100 kc. The period of the oscillator output is, therefore, 10 microseconds. The output signal from this crystal oscillator is developed across coil *L1* of tertiary-wound transformer *T103* and is inductively coupled to coil *L2*. Coils *L2* and *L3* are part of the blocking-oscillator circuit of tube *V109A*, having a pulse repetition rate of 10 microseconds. Resistor *R152* and capacitor *C144* act as the time-constant arrangement for this blocking oscillator. Negative 10-microsecond pulses are available across plate-load resistor *R148* as markers. Positive 10-microsecond marker pulses are available across cathode-load resistor *R149* of tube *V109A*.

Tube *V110A* acts as another blocking oscillator. Transformer *T104* is exactly the same as *T103*. In *T104*, however, two of the coils are series-aiding connected in the plate circuit of the blocking oscillator; the third coil is in the cathode circuit, instead of the grid circuit as in the normal blocking oscillator. The circuit nevertheless functions in a similar fashion.

This circuit is set into oscillation by the 10-microsecond positive-pulse output from the first blocking oscillator. This positive pulse is coupled to the grid of *V110A* via coupling capacitor *C145*. Grid resistors *R150* and *R151*, in conjunction with capacitor *C145*, form the time-constant network for this second blocking oscillator. Resistor *R150* is made variable so that the tube can conduct on either the fourth, fifth, or sixth 10-microsecond marker. This control is normally adjusted so that the oscillator fires on every fifth 10-microsecond marker, thereby having the blocking oscillator produce 50-microsecond pulse markers. As in the first blocking-oscillator circuit, positive 50-microsecond markers are obtained across cathode resistor *R154* and negative 50-microsecond markers are obtained across plate-load resistor *R153*.

In this particular instrument, the markers are fed to the cathode of the cathode-ray tube. Either positive or negative 10- or 50-microsecond pulse markers can be selected. The strength of the positive markers is such that, when applied to the cathode, the effective bias on the grid exceeds cutoff. Consequently, the positive pulses cause blanking markers equally distributed along the pattern being observed. The negative markers, when applied to the cathode, effectively appear at the grid as a positive signal and hence increase the intensity of the beam. Therefore, when negative marker pulses are chosen, they will appear as bright spots equally distributed along the pattern.

One may question the use of blocking oscillators as the horizontal sweep in television receivers if the resulting sawtooth is not very linear. The nonlinearity of these resulting sawtooth waves occurs, as mentioned before, at the beginning of the rising portion of the curve. This can be seen in the photo of Fig. 9-80C. This degree of nonlinearity would produce a defect in the resultant picture if that portion of the sawtooth wave were active in reproducing the picture. In television, however, part of the upper and *lower* portion of the rising part of the sawtooth signal is "blanked" out along with the retrace section of the signal. This means that only the linear part of the signal is effective in scanning the picture.

Single-Pentode Relaxation Oscillator

The sweep oscillators described thus far are the most common types used today as linear sawtooth generators for oscilloscopes and television receivers. There are numerous other types of linear sweep oscillators, but they are very seldom employed in oscilloscopes or television receivers. A number of these sweep circuits still fall under the broad classification of relaxation oscillators, because they use the charging and discharging of a capacitor through a resistance to produce the sawtooth wave. It is beyond the scope of this book to cover each one of these circuits in detail. A few of these circuits which are well known and more important will be briefly considered here. For further detailed information regarding these sweep circuits as well as other types, it is suggested that the interested reader consult the bibliography for this chapter.

About the same time, H. J. Reich, of the United States, and B. C. Fleming-Williams, of England, developed a one-tube relaxation oscillator employing a pentode.[11, 12] The operation of both circuits depends upon a variation in suppressor voltage on the tube. Both circuits are critical in operation in that their action in producing the sawtooth wave depends upon sudden changes of plate current when the plate voltage reaches certain critical values. The circuit developed by Fleming-Williams appears in Fig. 9-87A. Let us briefly analyze this circuit first.

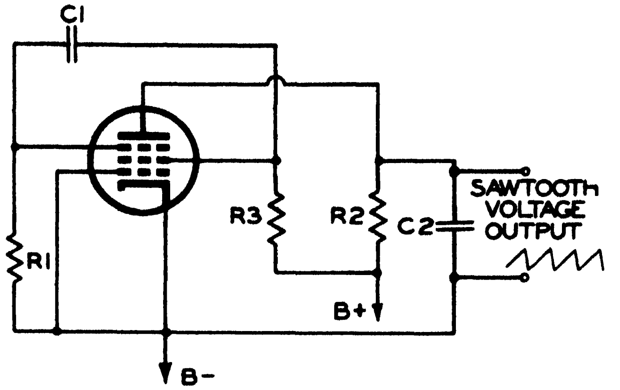


Fig. 9-87A.—Circuit of a single-pentode type relaxation oscillator developed by Fleming-Williams.

Assume a starting point with capacitor *C2* in an uncharged condition (i.e. no voltage across it) and the suppressor grid biased negative so that there is negligible plate-current flow. Potential applied to the screen grid causes practically all of the electrons emitted from the cathode to flow in the screen-grid circuit. The screen-grid current flowing through *R3* reduces the screen potential, which, in turn, means a further reduction of the suppressor-grid voltage, because the screen and suppressor are coupled together through *C1;* the plate current remains practically zero. While this is occurring, capacitor *C2* begins to charge up from the B− supply through the plate-load resistor *R2*. As *C2* charges, the potential on the plate of the tube increases. The charging of capacitor *C2* represents the linear rising part of the sawtooth output signal. A point will be reached where the voltage on the plate of the tube will be high enough to start drawing plate current through the negatively biased suppressor grid.

This flow of plate current reduces the flow of screen current because, before the plate drew current, all the cathode current flowed to the screen. With the screen current being reduced, the voltage at the screen rises. This rise in voltage is coupled to the suppressor grid via *C1* and results in the negative bias on this grid being reduced. This in turn increases the plate current still more, which means a further reduction in screen current, a rise in screen voltage, further increase in suppressor potential (in the positive direction), and an additional rise in plate current. This action is cumulative. During this period of reduction in screen current, capacitor *C1*, which is in a charged state, discharges through *R1* and *R3*. The discharge of *C1* causes the suppessor to be driven in the positive direction. By having a variation in suppressor voltage, the plate-voltage plate-current characteristic curve of the tube is such that a point will be reached where the plate current suddenly increases very rapidly. This rapid increase in plate current greatly reduces the effective plate resistance of the tube, and capacitor *C2* discharges through the tube. This discharge is very rapid because the dynamic curve is quite steep at this point, and in a short period of time the plate current is again reduced. This discharge of *C2* represents the flyback or retrace of the sawtooth wave.

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[11] Reich, H. J., "Trigger circuits," *Electronics,* vol. 12, pp. 14-17, August, 1940.

[12] Fleming-Williams, B. C., "Single-Valve Time-Base Circuit," *Wireless Eng.*, vol. 17, pp. 161-163, April, 1940.

As soon as the plate current is reduced, the screen current begins to increase again and the screen voltage begins to drop. This drop is coupled to the suppressor grid via *C1,* and hence the suppressor grid becomes less positive. This further reduces the plate current and hence increases the screen current. This action of the plate current being reduced and suppressor grid going negative is cumulative, and a point will be reached where the plate current is practically reduced to zero and capacitor *C2* starts to charge up again. We are now back again to the point where we started the cycle of operation.

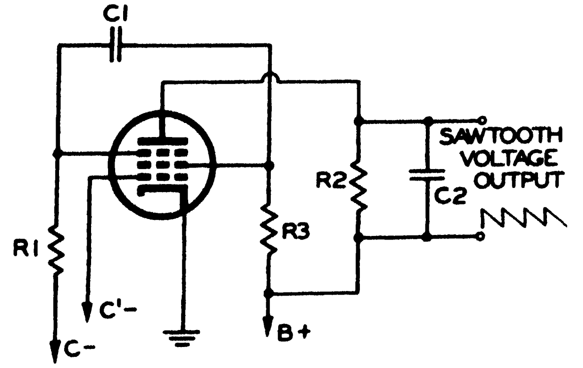

Fig. 9-87B.—A similar circuit to that in Fig. 9-87A, developed by H. J. Reich.

The circuit developed by H. J. Reich appears in Fig. 9-87B. In this circuit, the output capacitor *C2* has one end connected to B+, whereas in the circuit of Fig. 9-87A it is connected to B−. Also note that the suppressor and control grids are connected to fixed negative supply sources. Let us assume a starting condition where the total plate-supply voltage is applied to the plate of the tube. Plate current starts to flow at its maximum value and charges up capacitor *C2.* As the capacitor becomes charged, the plate current decreases. This decrease in plate current causes an increase in screen current and hence a decrease in screen voltage. This decrease in screen voltage is coupled to the suppressor grid via *C1* and drives this grid more negative than it was before, thereby further reducing the plate current. This action is cumulative.

As capacitor *C2* charges, due to plate-current flow, the plate voltage will decrease to a certain value, often called the "critical" value, where the dynamic characteristic curve of the tube is such that the plate current is virtually reduced to zero. When this happens, the tube is effectively cut off and capacitor *C2* starts discharging through *R2.* As *C2* discharges, the plate voltage starts to rise again, and after a period of time maximum plate current starts flowing. At this point, the screen current drops, the screen voltage is increased, and the suppressor grid is driven in the positive direction. The cycle of operation starts over again, with capacitor *C2* charging up due to plate-current flow.

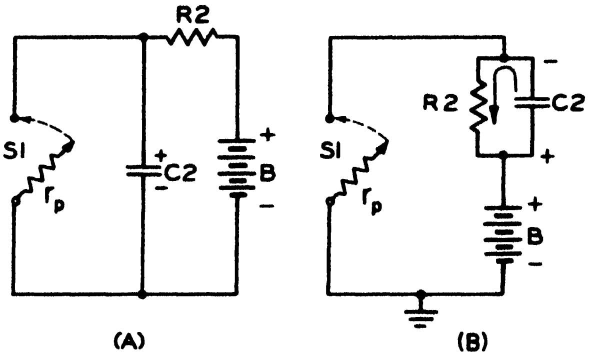

Fig. 9-88.—Equivalent (simplified) circuits for the single-tube relaxation oscillators of Fig. 9-87 are shown in (A) and (B), respectively.

The output sawtooth waveform from the circuit of Fig. 9-87B is different from that of Fig. 9-87A, in that the charging of *C2* represents the flyback or retrace of the sawtooth wave which uses the low plate resistance of the tube to complete the charging circuit. The discharge of *C2* through *R2* represents the linear section of the sawtooth wave. Thus the charge and discharge time of the sawtooth in the Reich circuit is the reverse of that in the Fleming-Williams circuit. The primary difference between these two circuits is easily seen from the two equivalent circuits of Fig. 9-88. Part (A) is for the circuit of Fig. 9-87A and (B) for the circuit of Fig. 9-87B. Switch *S1* in each equivalent circuit is inserted to indicate the transfer from the conduction to the nonconduction period of the tube (i.e. when the plate current is practically zero). With the switch open, the tube is in the nonconducting state. Resistor r_p in the drawings indicates the low value of plate resistance of the tube; when the switch is closed, this resistance is in the conducting circuit.

Equivalent circuit (A) shows that capacitor *C2* charges up from the B− supply through *R2* during the nonconduction period of the tube, and discharges through the plate resistance of the tube during its conducting period. The charging period of *C2* in this case is the linear rising portion of the sawtooth output waveform, and the discharge is the flyback time of the signal. The charge period is quite long compared to the discharge period. In this circuit, the tube acts as a shunt switch to the capacitor. In equivalent circuit (B), the tube acts as a switch in series with the capacitor. In this latter circuit, when the switch is closed (i.e. conduction period of the tube), *C2* charges up due to plate-current flow; when it is open, *C2* discharges through *R2* in the direction indicated. The discharging period for (B) represents the linear *decaying* section of the sawtooth output signal and the charging period is the retrace time of the wave which has a *rising* characteristic. The discharge period is quite long compared to the charge period. A high value of *R2* will effectively reduce the charging period of *C2* and thus provide a quicker retrace time per cycle.

ELECTROMAGNETIC SWEEPS

In the discussion of sweep circuits thus far, we constantly referred to a linear sweep as the necessary signal to produce the horizontal deflection when observing certain electrical phenomena that are applied to the vertical plates of a cathode-ray tube. Linear horizontal deflection is always necessary in order to reproduce an undistorted picture on the screen of the cathode-ray tube. A *sawtooth voltage* wave is the required sweep signal to be applied to the deflection plates of an electrostatic cathode-ray tube, but cannot be used directly as the sweep voltage for electromagnetic cathode-ray tubes. The output signal from all of the sweep circuits covered in this chapter so far was a sawtooth voltage, and as such could only be used for electrostatic cathode-ray tubes. However, this does not mean that these sweep circuits cannot be used for electromagnetic deflection. Only slight circuit modifications are required to make these circuits useful as electromagnetic sweeps.

Waveform Analysis In Deflection Coils

In Chapter 6, we mentioned that the *current* through the deflection coils, not necassarily the voltage applied across these coils, had to be linear for proper deflection of the electron beam. In order to have linear or sawtooth current flow through the deflection coils, the voltage applied across the coils must have a different waveshape. We have to know something about the reactive and resistive components of the deflection coils in order to find out what is the required shape of the voltage waveform across the coil.

Every coil has inductance, resistance, and distributed capacitance. The distributed capacitance of deflection coils as used in television receivers is generally too small to have any effect on the current flowing through the coil at the usual frequencies of operation. Only the inductance and resistance of the coil are considered as affecting the current waveform. These two components are not separated, but are evenly distributed throughout the coil. We can, however, represent the circuit of this coil by a series arrangement of a lumped resistance and inductance as a convenient means of illustration without introducing any errors. We know that the current flowing through this series *R-L* arrangement has to be linear, that is, have a sawtooth waveform. On the basis of this knowledge we can work backwards to find the required voltage waveform to be applied across the coil. In other words, the voltage waveforms required to produce a sawtooth of current through a pure inductance and resistance are found individually. The combination of these two voltage waveforms, then, will result in the required voltage waveform to be applied across the deflection coil.

First we will consider the resistive part of the circuit. A pure resistance will not cause any phase change between the voltage across it and the current flowing through it. Thus, if a sawtooth waveform of voltage is applied directly across the resistor, a sawtooth waveform of current will flow through it. This is basically indicated in part (A) of Fig. 9-89. The amplitude or strength of the current flow is determined directly by Ohm's law, $I = E/R$, where E is the applied voltage, R the value of resistance, and I the current.

The inductive part of the deflection coil presents the interesting problem. It is a known fact of theory and experimentation that when a steadily changing current (that is, one having a linear rise and fall similar to a sawtooth) passes through an inductance, a fixed steady voltage is produced across this coil for each rise and fall of current. In other words, the voltage is proportional to the rate at which the current is changing and also to the size of the inductance. Such a voltage would have a rectangular waveshape as indicated in part (B) of Fig. 9-89.

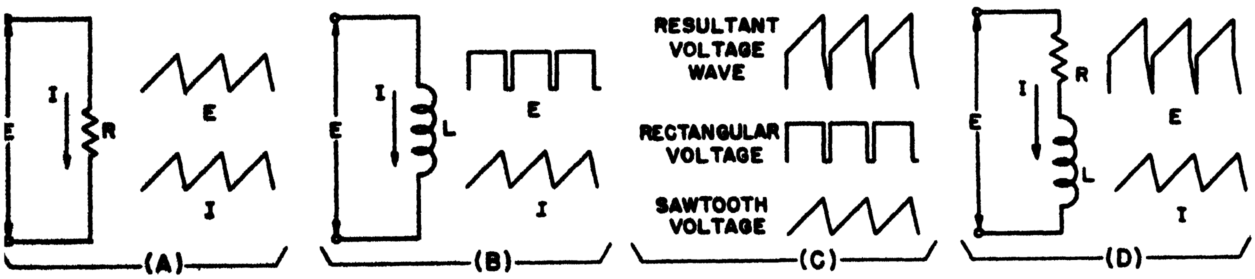

Fig. 9-89.—Graphical analysis of the modifications of waveforms produced by deflection coils.

Since our inductance here is fixed in value, any change in the voltage waveform across it must be due to a change in the current through it. And the current does change; first it rises fairly slowly; that is, it changes in the positive direction. As a result, there is a positive voltage across the inductance which endures for exactly the same length of time as the current rises. Then, abruptly, the current starts to decrease; that is, it changes in the negative direction (this occurs during the retrace of the sawtooth current wave). At the time that this change in the current wave takes place, there is a corresponding change in the voltage wave; and, since the current is changing steadily in the negative direction, the voltage is steady at a negative value. The duration of the negative swing of the voltage is exactly the same as the duration of the negative change of current that produces it. Thus, alternating fixed positive and negative voltages are produced by the passage of a sawtooth wave of current through an inductance.

Let us now see what the resultant waveshape is when the sawtooth voltage at (A) and the rectangular voltage at (B), of Fig. 9-89, are combined. Since both voltages produce the same current waveshape, the retrace of the sawtooth voltage occurs exactly at the same time as that of the negative going part of the rectangular signal. The result of combining two such voltages appears in Fig. 9-89C. The peak-to-peak amplitudes of the sawtooth and rectangular waveshapes in this figure are drawn equal for the sake of illustration. In part (D), the equivalent *R-L* circuit of the yoke is shown in connection with the resultant voltage waveform required to produce a sawtooth of current through the coil.

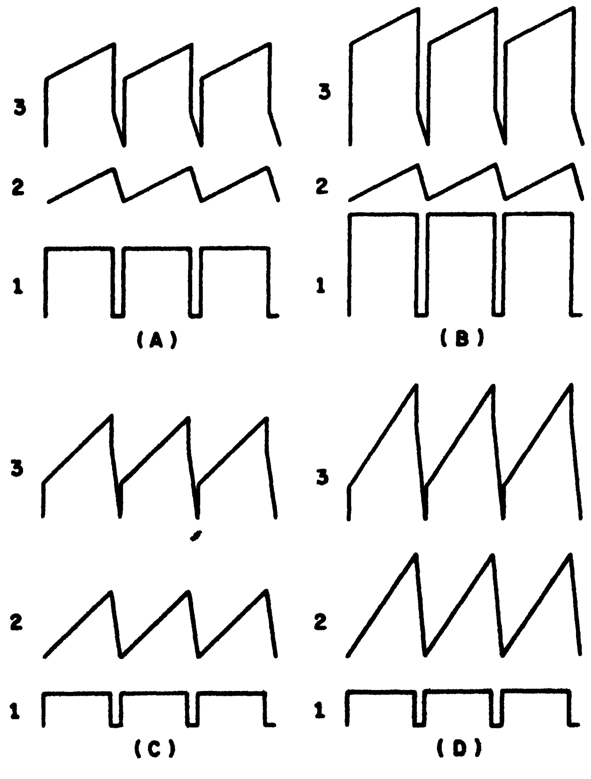

Fig. 9-90.—Resultant voltage waveforms required for deflection coils, when the voltage drop across the resistance and inductance are not equal.

In practice, the resistive and inductive components of the deflection coils will vary somewhat. If the resistive component of the coil is high compared to the inductive reactance, then the resultant voltage across the coil should be closer in shape to its sawtooth component. This is so because the resistive component of voltage will be higher. On the other hand, if the inductive reactance is higher, the resultant voltage will be closer in shape to its rectangular component because the voltage drop across the inductance will be higher. In Fig. 9-90, we have illustrated numerous resultant voltage waveforms that are required when the voltage drop across the resistance and inductance are not equal. The case of equal component voltages was shown in Fig. 9-89C.

In all parts of Fig. 9-90, the rectangular waveshape at *1* represents the voltage drop across the inductance, that at *2* the voltage drop across the resistance, and that at *3* the waveshape of the voltage across the series combination of the inductance and resistance. The resultant waveforms at parts (A) and (B) indicate two cases where the inductive reactance is greater than the resistance. For the waves at (A), the inductive reactance is twice as great as the resistance, and at (B) it is three times as great. The resultant waveshapes at (C) and (D) occur when the resistance is greater than the inductive reactance. For the wave at (C), the resistance is twice as great as the reactance, and at (D) it is three times as great. Note that in parts (B) and (D) the resultant waveshapes appear more like the higher voltage component of which it is composed. Resultant waveforms of the type shown are generally identified as *trapezoidal* in shape. The term is derived from the fact that the leading and trailing edges of the wave are roughly parallel, while the top portion and base are not.

Obtaining the Trapezoidal Waveform

Now that we have determined the shape of the voltage waveform that must exist across the coil in order to achieve the desired sawtooth deflection, our next logical step is to find out how to produce such a waveform. There are a number of methods which are used today, but the most common ones employ unsymmetrical multivibrators or blocking oscillators similar to

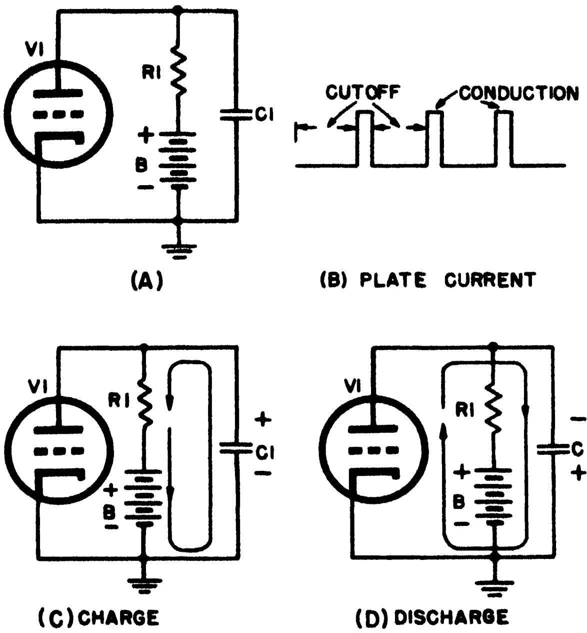

Fig. 9-91.—A circuit required for producing a sawtooth voltage wave is shown at (A), as well as the shape of the plate-current pulse (B). The directions of current flow for charge and discharge are shown at (C) and (D), respectively.

those described previously. When these circuits are employed to produce a sawtooth of voltage, a capacitor is placed across the output. This is basically illustrated in Fig. 9-91A. In this circuit, *V1* represents the output tube, *R1* the plate-load resistor, and *C1* the capacitor employed to yield the sawtooth wave for electrostatic deflection.

The production of a sawtooth wave with a sharp retrace is dependent upon the output-tube plate current flowing in short pulses. In other words, the tube should be conductive for a much shorter period of time than it is nonconductive. This means that the plate-current waveform is ideally in the shape of a rectangular wave as shown by the drawing at (B). When the tube is not conducting, capacitor *C1* charges up from the B— supply through resistor *R1*, and during the short periods of conduction, *C1* discharges through the tube. The direction of electron flow for the charge and discharge of *C1* is illustrated in Figs. 9-91C and D, respectively. From these two drawings, it is seen that the currents flow in opposite directions. During the charge of *C1*, the plate side of the capacitor is positive with respect to its other side, and the reverse during its discharge.

We do know that with the circuit as it stands now, a sawtooth voltage appears across this capacitor. If we combine this sawtooth voltage with a rectangular wave, we can obtain the trapezoidal waveform needed for electromagnetic deflection. Inserting a resistor in series with the capacitor of the circuit of Fig. 9-91A will help produce the desired waveform. The circuit for this latter arrangement is shown in Fig. 9-92, with *R2* being the added resistor. In order to produce this trapezoidal waveform, the voltage appearing across resistor *R2* should have a leading and trailing edge that are both vertical. If both vertical sides are approximately equal, the voltage across the resistor will appear as a square wave, as ideally indicated in the drawing. The value of *R2* is usually made much smaller than *R1* so that it will have very little effect on the total charging and discharging time. As far as capacitor *C1* is concerned, the voltage appearing across it will still have the same sawtooth waveform. With the circuit arranged as shown in Fig. 9-92, the charging and discharging currents both flow through resistor *R2*. Because of this fact, it may, at first, be difficult to conceive how the necessary voltage waveshape across this resistor comes about. Let us now analyze the formation of the voltage waveshape across *R2*.

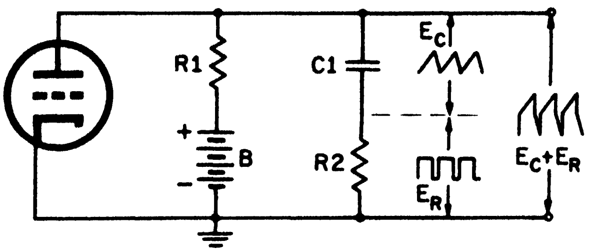

Fig. 9-92.—A modified version of the circuit shown in Fig. 9-91A is shown here, where a resistor *R2* has been inserted in series with the capacitor.

During the period of charge, when the tube is cut off, the initial charging current is a maximum and decreases exponentially. Since *R2* is in the charging path, and also since the voltage across a resistor is in phase with the current flowing through it, the voltage across *R2* will rise suddenly to a maximum from point *a* to *b* as seen in Fig. 9-93A. The time constant of charge is sufficiently long so that the exponentially decreasing charging current and hence voltage across *R1* will decrease only slightly by the time the tube starts conducting. This is indicated in Fig. 9-93A from time *b* to *c*. During this period of charge, the voltage across *R2* is positive with respect to the ground because of the direction of current flow.

The moment the tube becomes conductive, *C1* stops charging and starts discharging. The initial discharge current rises instantaneously to a maximum; hence a maximum voltage drop exists across *R2*. This is indicated in Fig. 9-93B by the instantaneous voltage rise from *c* to *d*. As the capacitor continues to discharge exponentially, the voltage drop across *R2* decreases.

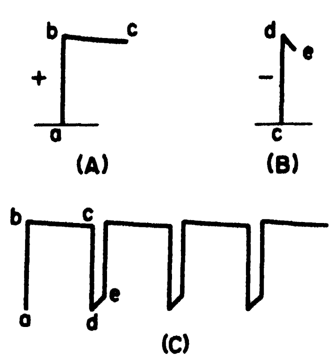

Fig. 9-93.—Voltage charge and discharge curves for the circuit of Fig. 9-92 are shown in (A) and (B), respectively, while the resultant combined waveshape is shown in (C).

The conducting period of the tube is much shorter than the cutoff period and the time constant of discharge much smaller than that of charge, so that the voltage across *R2* decreases as shown during time *d* to *e*.

Since the discharge current flows in the opposite direction to the charging current, the voltage drop across *R2* during discharge is considered negative with respect to that during charge, and is so indicated in the drawing. Combining the voltage drops across *R2* during charge and discharge, in their proper polarity, to represent a continuous waveform, results in the curve shown in Fig. 9-93C. From this latter drawing, one can readily see how the waveform across resistor *R2* in Fig. 9-92 approximates a square wave.

It should be remembered that the curves drawn in Figs. 9-89 to 9-93 are idealized, and in actual practice slight modifications occur. In order to illustrate some actual waveforms, a number of oscillograms were taken from the output of a plate-coupled multivibrator. The circuit used was similar to that of Fig. 9-57, except that a resistor was inserted between the capacitor *C3* and ground. Thus we can assume that the output circuit of Fig. 9-57 relative to tube *V2* can now be represented by the circuit of Fig. 9-92. The waveforms observed at different points in this system are indicated in Fig. 9-94.

The oscillogram at (A) was taken at the plate of the output tube with the capacitor and resistor disconnected from the circuit. This photo indicates the rectangular voltage output from the plate circuit of the asymmetrical multivibrator. With the capacitor and resistor inserted into the circuit, the voltage waveform at the output is indicated by the photo at (B). Note how closely this oscillogram compares with the line drawings illustrated in Figs. 9-89C, 9-89D, 9-90, and 9-92. The voltage waveform across the capacitor alone appears in part (C), and that across the resistor alone is indicated in part (D). Note how closely the photo at (D) resembles the theoretical drawing of Fig. 9-93C. The oscilloscope used for the observation of the waveforms of Fig. 9-94 was adjusted to give the best possible picture for illustrative purposes; consequently, the amplitudes of the waveforms are not necessarily in proper proportion to each other.

Blocking oscillators are used a great deal in television receivers as sweep circuits for electromagnetic deflection. One method is to employ a circuit similar to that shown in Fig. 9-79, with the only modification being the insertion of a resistor in series with capacitor *C2*. With this addition, the output network of tube *V2* can also be represented by the simple circuit of Fig. 9-92.

It should be remembered that the plate current of the blocking oscillator flows only in pulses of short time duration and that these pulses are not rectangular or

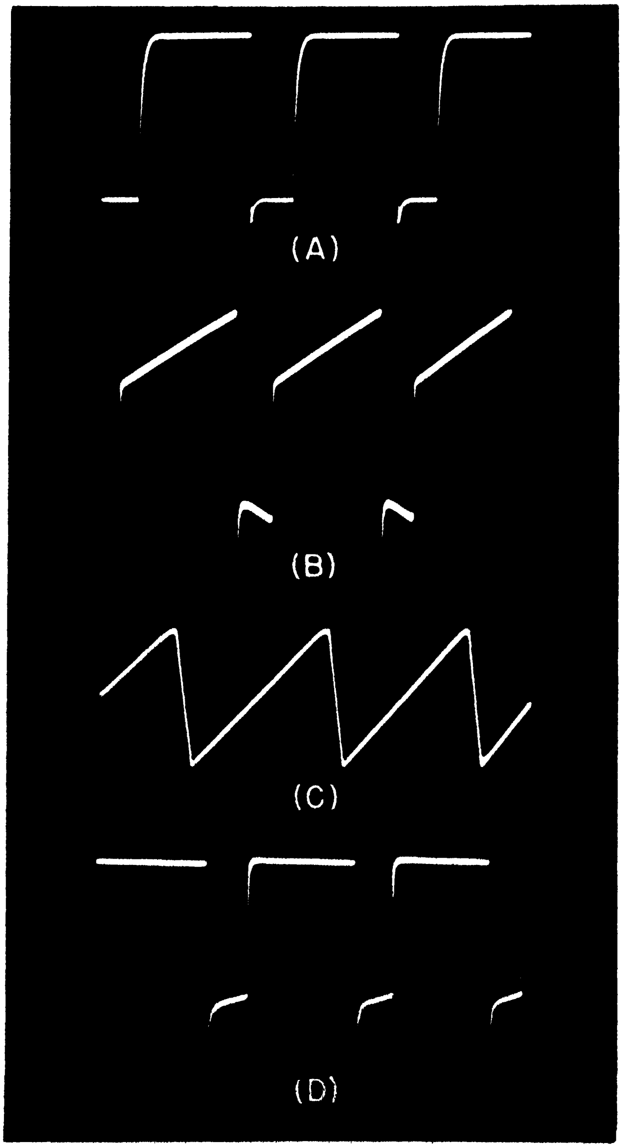

Fig. 9-94.—Actual oscillograms of waveshapes produced by a circuit similar to that shown in Fig. 9-92.

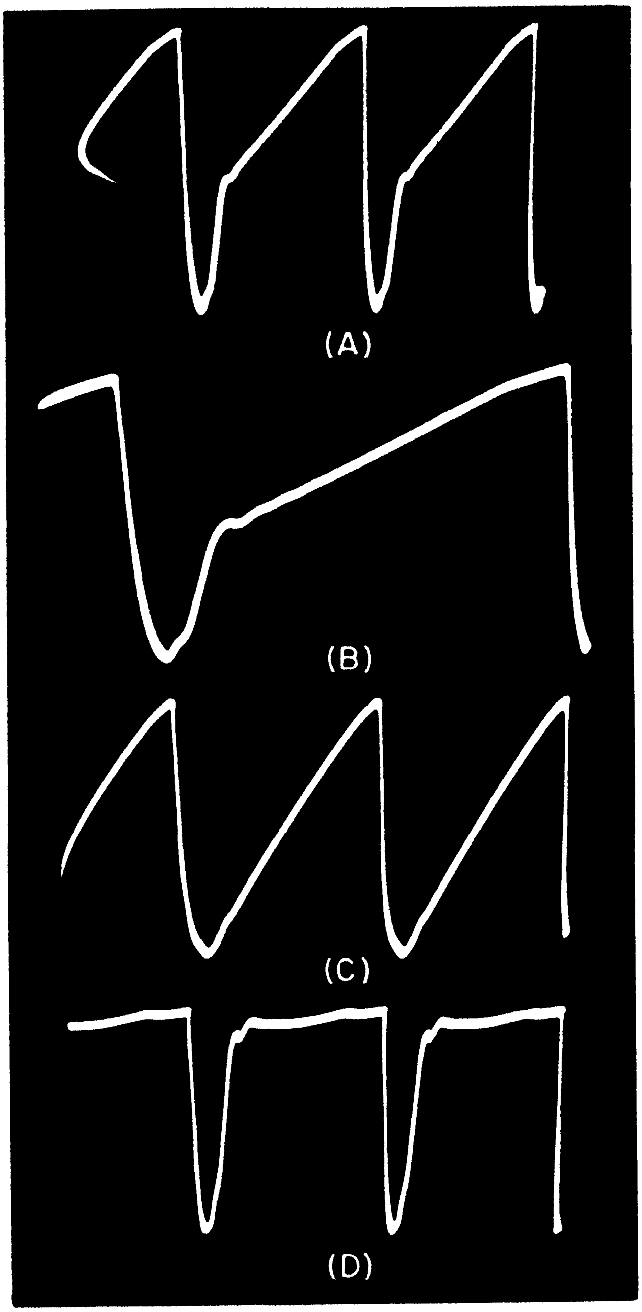

Fig. 9-95.—Oscillograms of the blocking oscillator shown in Fig. 9-79, after modification for electromagnetic sweep.

sinusoidal due to the nature of the circuit. The exact shapes of these pulses were already indicated by the oscillograms of Fig. 9-84. Therefore, it would be difficult to obtain a waveshape exactly the same as Fig. 9-94B from the output of the modified blocking oscillator. The oscillograms of Fig. 9-95, which were taken from the blocking oscillator circuit of Fig. 9-79 modified for electromagnetic sweep, prove this point. The photo at (A) of Fig. 9-95 indicates the waveshape of the output of the series *R*-*C* network. Note how the shape of this signal differs from that of Fig. 9-94B. One cycle of this signal spread out for a clearer indication of the waveshape is shown in part (B) of Fig. 9-95.

The waveshapes of the signal voltage across the capacitor and resistor are indicated in parts (C) and (D), respectively. Note how the signal across the resistor deviates from a rectangular waveform. The shape of the negative peaks of this signal is due to the nonrectangular form of plate-current pulses, and the small transient ripple on the top or positive portion of the wave is primarily due to oscillation of the blocking-oscillator transformer. The sawtooth of voltage across the capacitor, part (C), does not, however, appear much different from a true sawtooth. The primary difference in the waveshape is at the bottom of the retrace and the start of the trace. This section of the sawtooth sweep is blanked out in television receivers, as mentioned previously, and, therefore, cannot do any harm to the reproduced picture.

Hartley Oscillator Sweep Circuit

A common type of electromagnetic sweep circuit employed in television receivers for the horizontal deflection is illustrated in Fig. 9-96. Pentode *V1* is designed to operate as an electron-coupled Hartley sine-wave oscillator. The cathode, control grid, and screen grid function as a normal triode oscillator, where *L* and *C* are the tank-circuit components which determine the frequency of oscillation. Inductance *L* is usually preset to produce the desired horizontal frequency. Components *C1, R1,* and *R2* are the time constant and grid-bias networks for the oscillator tube. Resistor *R2* is made variable so that it can control the frequency of operation of the oscillator over a small range of frequencies.

With the circuit operating, a sine wave of voltage appears in the oscillator grid circuit and is electron-coupled to the plate of the tube. During part of the cycle of oscillation, the bias developed on the grid of tube *V1* is great enough to cause plate-current cutoff. This means that the signal appearing at the plate of the tube is not a pure sine wave, but is unsymmetrically clipped on its positive half-cycles. This distortion is a requirement for the proper operation of the system. The signal from the plate of *V1* is passed through a resistor-capacitor differentiating network consisting of *R3*-*C3* and *R4*-*C4* for pulse-shaping purposes.[13] In other words, the clipped sine-wave output from *V1* is fur-

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[13] Differentiating as well as integrating networks are discussed in most good texts on this theory.

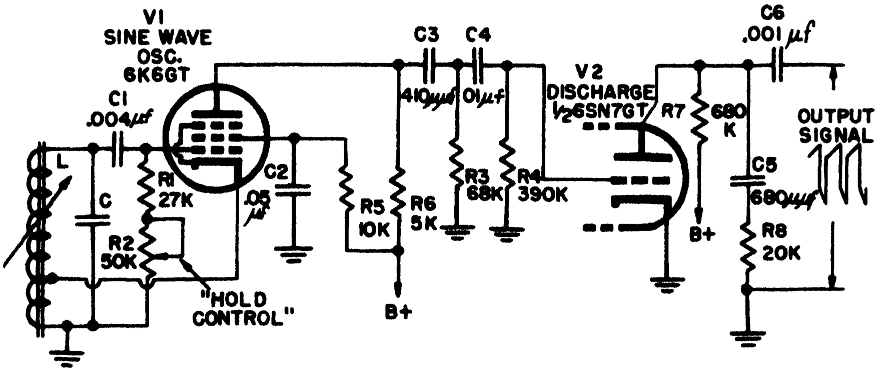

Fig. 9-96.—A conventional type of electromagnetic sweep circuit used in television receivers for horizontal deflection.

ther distorted by the differentiating network to produce a signal which appears like a sine wave with positive pulses added to its positive half-cycles. The signal output from the differentiating network is coupled to the grid circuit of tube *V2*.

This circuit consisting of tube *V2* and components *C5* and *R8* is employed to produce the final voltage waveshape for electromagnetic deflection. The network is very similar to those discussed previously, in that it utilizes the charge and discharge of capacitor *C5* to help produce the desired waveshape. Conduction of tube *V2* occurs only during the time that the positive pulses appear on its grid and is in a state of nonconduction or cutoff during the remainder of the cycle.

Capacitor *C5* charges up from the B supply through resistor *R7* during the cutoff periods of tube *V2*. When the pulses are applied to the tube, conduction occurs and *C5* discharges through the low plate resistance of *V2*. The charge and discharge current of *C5* is shaped roughly like a rectangular wave. This results in a sawtooth voltage across *C5*, a rectangular voltage across *R8*, and across their combination we have the required trapezoidal voltage waveshape for electromagnetic deflection, as seen in Fig. 9-96. Capacitor *C6* is used merely to couple the final signal output from the discharge tube to the horizontal-output tube of the television receiver.

The values indicated next to each component are those required to operate the circuit at 15,750 cps, the horizontal sweep frequency of a television receiver. Although not indicated in the drawing, this type of sweep circuit usually has the output from a sync-discriminator transformer coupled to the tank coil *L* of the sine-wave oscillator. Also, a reactance-tube circuit is usually capacitor-coupled to the oscillator. The discriminator, in conjunction with the reactance tube, serves the purpose of keeping the frequency of the oscillator in exact step with that of the horizontal scanning frequency at the transmitter.

Many circuits that can produce a pulse whose frequency is fairly accurate may be used as the basis of a sweep circuit. Once the pulse is generated, all that is needed is a simple discharge-tube circuit similar to that illustrated in Fig. 9-96. If straight linear electrostatic deflection is required, the resistor in the output circuit of the discharge tube is omitted. On the basis of this simple pulse requirement, numerous other networks can be used as sweep circuits. It is beyond the scope of this text to discuss them all.

LINEARIZING THE SAWTOOTH TRACE[14-16]

In all the time bases discussed thus far, the charging and discharging of a capacitor was put to use in establishing the sweep voltage. For electrostatic systems, the necessary sweep voltage has a sawtooth waveform which is taken across a capacitor in the output of the sweep circuit in question. For electromagnetic systems, the majority of sweep circuits make use of a combination sawtooth signal and rectangular wave to produce the required sweep-voltage waveform. The sawtooth signal in both cases is obtained as a result of capacitor charge and discharge. Thus we see that a sawtooth signal as produced by the charge and discharge of a capacitor is the sweep voltage for electrostatic deflection and part of the sweep signal for electromagnetic deflection.

If the rising portion of the sawtooth signal, that is, the trace, is nonlinear, then the reproduced signal on the screen of the cathode-ray tube will be distorted. This distortion is usually manifested by a bunching of the picture at the right end. This was indicated in the

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[14] Clarke, A. C., "Linearity circuits," *Wireless Eng.*, vol. 21, pp. 256-266, June 1944.

[15] Owen, R. P., "Linear sweep circuits," *Electronics*, vol. 19, pp. 136-138, December 1946.

[16] Puckle, O. S., "Time Bases," 1st edition, John Wiley & Sons, Inc., New York, N. Y., 1943; Chapter VII.

oscillograms of Fig. 9-18 where a nonlinear sawtooth signal at (A) reproduces a distorted wave shown at (B). Part (C) of this figure graphically illustrates the unequal movement of the spot across the screen for equal time intervals along a nonlinear sawtooth sweep. A complete graphical illustration, showing how a nonlinear sawtooth sweep produces a distorted sine wave, is shown in Fig. 9-22.

The voltage charge and discharge across a capacitor is an exponential one; in most of the circuits analyzed in this chapter, it is a rising curve, while the discharge is a decaying one. The trace of the sawtooth signal is usually obtained from the charge part of the cycle and the retrace from the discharge part. Since the rise time is exponential, the trace of the sawtooth signal can never be perfectly linear. In other words, the trace cannot be the ideal straight line that we desire. Since at the beginning of an exponentially rising curve, the slope does not change very much, the curve at this point will be most linear. Consequently, a sawtooth signal whose maximum voltage is only a small portion of the total available charging voltage on the capacitor will be most linear.

Quite a number of different circuits exist today for improving the linearity of a sawtooth sweep, the majority of them being patented in the United States and Great Britain. These linearizing circuits can be classified into one of the following principal groups:

1. Constant-current devices
2. Additional time-constant circuits
3. Feedback circuits
4. Nonlinear amplifiers.

Some sweep circuits may employ more than one method of linearization. This is especially true in television receivers where the effect of nonlinearity is more pronounced than in oscilloscopes. In this section, we will only discuss those circuits that are used most often; for further references, consult the bibliography.

Constant-Current Pentode

In most types of sweep circuits, the capacitor across which the sawtooth wave is produced charges up from the B supply through a resistor. This resistor limits the flow of charging current, which does not flow at a constant rate but decreases in exponential fashion. If this charging resistor can be replaced by a device which will limit the flow of charging current at a constant rate, it is possible to obtain a fairly linear sawtooth wave. One of the first circuits to make use of this idea was the popular thyratron relaxation oscillator.

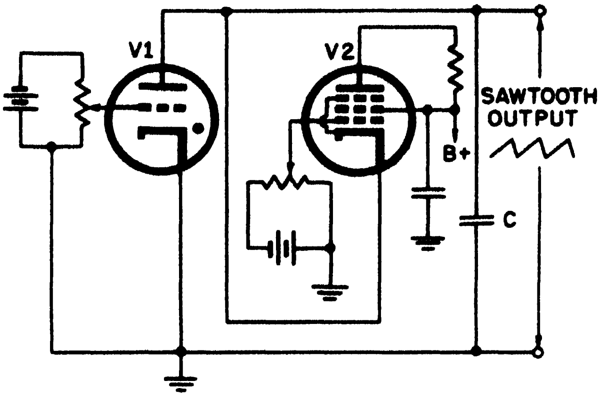

Fig. 9-97.—A thyratron relaxation oscillator circuit using a pentode as constant-current device.

In this new type of thyratron circuit, a pentode is normally used as the constant-current device taking the place of the resistor. A simplified form of such a circuit is illustrated in Fig. 9-97, where tube *V2* is the pentode that takes the place of the charging resistor. The pentode is designed to operate over a range where the plate current is fairly constant. When plate voltage is first applied to the circuit, capacitor *C* charges up from the B supply through tube *V2*. As the plate current flowing through *V2* charges capacitor *C*, the voltage charge across the capacitor increases. Since the plate current of tube *V2* remains fairly constant, the increase in voltage across capacitor *C* will increase at a constant rate and hence has a linear rise.

Discharge of *C* occurs when the voltage across it reaches the ionizing potential of the thyratron tube *V1*. The frequency of operation of such a circuit is primarily dependent upon the value of capacitor *C* and the amplitude of the charging current, that is, the current flowing through tube *V2*. Variation of the charging current has the same effect as changing the value of resistor *R* in Figs. 9-34 or 9-35. The strength of the charging current can be varied easily by changing the bias on the tube. In the drawing of Fig. 9-97, this variation in bias is obtained from a potentiometer across a battery. When cathode bias is employed on the pentode, a potentiometer in the cathode circuit regulating the bias is normally the fine frequency control of the circuit, and the variation in the value of *C* is the coarse frequency control.

Today, few thyratron sweep circuits as used in oscilloscopes employ the constant-current pentode. When such a pentode is employed, it is used more often with a hard-tube sweep circuit, especially the multivibrator.

In actual practice, the pentode tube is used to replace the resistor through which the capacitor discharges rather than that through which it charges. Such circuits where the pentode is used as the discharge device are of simpler design and give somewhat better performance. A simplified circuit of this type appears in Fig. 9-98. In this drawing, *V1* represents a tube in which the plate current flows in pulses of short time duration. (This current flow is indicated by the rectangular wave on the left side of the figure.) In other words, the signal voltage applied to the grid of *V1* is such that the tube is cut off for much longer periods than it is allowed to conduct. The circuit functions as follows:

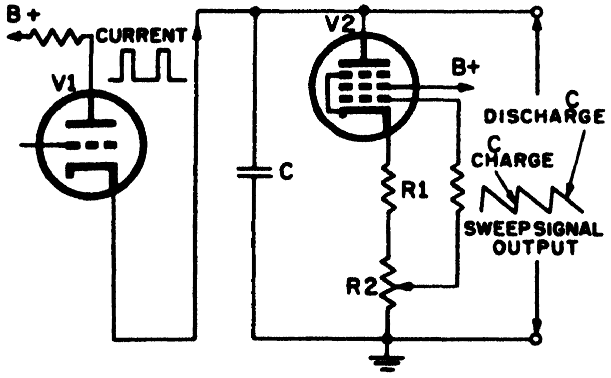

Fig. 9-98.—A sweep circuit using a pentode tube as discharge device.

During the short intervals of *V1* conduction, capacitor *C* charges up very rapidly from the plate-current flow of tube *V1*. Although the charging period represents the rise time of the resultant sweep signal, it is so rapid that the rise time or charging period is used as the *retrace* time of the sweep. The value of the screen voltage at tube *V2* is so adjusted that when *V2* conducts, its plate current is fairly constant. The moment that *V1* stops conducting, capacitor *C* discharges through tube *V2*. The plate resistance of the pentode is quite high, so *C* discharges quite slowly through *V1*.

The time constant of charge, which is low, is determined primarily by the product of capacitor *C* and the plate resistance of *V1* in series with its plate-load resistor. The time constant of discharge is chiefly determined by the product of *C* and the plate resistance of *V2*, and is much larger than the time constant of charge. Consequently, the trace time of the sweep signal is a decaying voltage represented by the discharge of *C* through *V2*. Due to the constant-current flow within tube *V2*, the discharging voltage across *C* follows a linear path rather than an exponential one. The interesting thing about the signal output is that its waveshape is the reverse of the normal signal output from most sawtooth sweep generators. In other words, in the normal sawtooth signal the trace of the signal is caused by the capacitor charge and the retrace by the capacitor discharge, whereas in this circuit the opposite is true.

Additional Time-Constant Circuit[17]

One of the simplest and cheapest methods of linearizing a sawtooth sweep is to employ some form of additional time-constant circuit. In most cases, a resistance-capacitance network is employed. This new circuit is designed to operate in such a manner that it will offer a correction voltage to the usual exponential rise time of the sawtooth wave so that the resulting sawtooth will be linear. Numerous circuit arrangements are possible, but only two will be considered here. These two, however, will give us a fair idea of how such new time-constant circuits should operate.

Let us first study the circuits of Fig. 9-99. Part (A) of this figure is a simplified form of the usual type of sweep circuit, where *V1* represents the tube in which the plate current flows in pulses of short time duration. Capacitor *C* charges up from the B supply through *R1* during the periods that *V1* is cut off; during the periods of *V1* conduction, *C* discharges through the tube. The sawtooth output is taken across *C* and may not be linear due to its exponential shape. Fig. 9-99B is the new circuit for improving the linearity of the sawtooth, where resistor *R2* and capacitor *C2* are the additional circuit elements.[18] Capacitor *C* of part (A) is effectively split in two and indicated as *C1* and *C3* in circuit (B). Capacitors *C1* and *C2* are approximately the same value, and resistor *R2* is about 0.5 megohm. The resultant sawtooth output is taken across capacitors *C2* and *C3* in series. Let us examine the operation of this circuit to see how it makes the sawtooth more linear.

Assume a starting point with all the capacitors charged up from the power supply. Capacitors *C1* and *C3* charge up through resistor *R1*, and capacitor *C2* charges up through resistors *R1* and *R2*. When *V1* starts conducting, all the capacitors start to discharge through the tube. Capacitors *C1* and *C3* discharge very rapidly, but *C2* discharges slowly because of the high value of *R2*. In other words, the time constant of discharge for capacitor *C2* is much higher than for the other capacitors. By the time *V1* becomes cut off, *C1* and *C3* are almost completely discharged, whereas capacitor *C2* has lost only a small quantity of its charge. *C1* and *C3* start to charge up again from the B

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[17] Keen, A. W., "Television time base linearisation," *Electronic Eng.* (London), vol. 21, pp. 195-198, June, 1949.

[18] "Television Topics—Linear saw-tooth oscillator," *Wireless World*, vol. 44, p. 425, May 4, 1939.

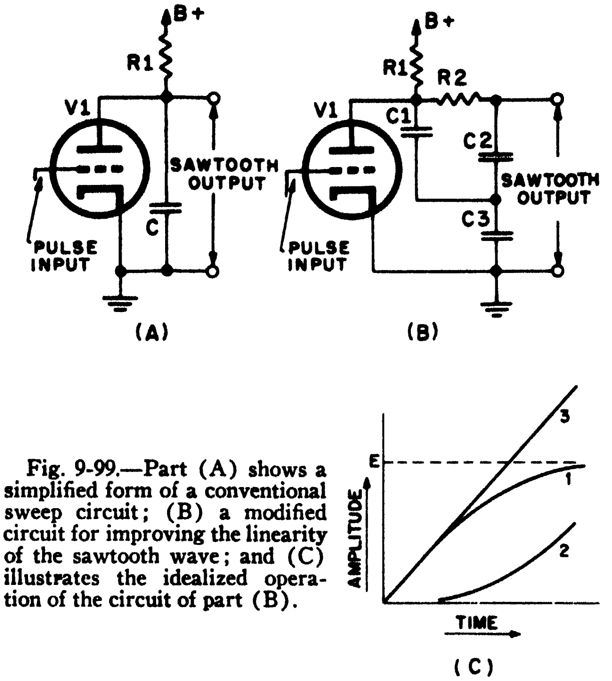

Fig. 9-99.—Part (A) shows a simplified form of a conventional sweep circuit; (B) a modified circuit for improving the linearity of the sawtooth wave; and (C) illustrates the idealized operation of the circuit of part (B).

supply. The voltage across *C2* is, however, still very high so that it continues to discharge through *R2* because the voltage across *C1* is very low. As *C2* continues to discharge, its electrons pile up on *C1*. Thus *C1* is charging up from two sources — from the B supply and also from *C2*. As the voltage across *C1* increases and that across *C2* decreases, a point will be reached where the voltages across each will be equal; *C2* will then stop discharging. Therefore, immediately after this period is reached, the voltage across *C1* will be greater than that across *C2*, and the latter capacitor will start to charge again.

From the circuit of Fig. 9-99B, we see that the output signal is taken across capacitors *C2* and *C3*. The voltage across capacitor *C3* is the nonlinear sawtooth with an exponentially rising characteristic. The voltage wave across *C2* is, however, much different. During part of the time during which capacitor *C3* charges, capacitor *C2* is discharging. The amount of discharging that *C2* undergoes results in a somewhat semicircular or parabolic waveshape across *C2*. Thus, while the exponential voltage across *C3* is convex, that across *C2* is concave. The time constants of the circuits are so arranged that the resulting waveform of the addition of the voltages across *C2* and *C3* is a fairly linear sawtooth.

The curves of part (C) ideally illustrate the operation of the circuit. In this graph, curve *1* is a normal exponential charging curve (where *E* is the maximum charging voltage) and, therefore, represents that of capacitor *C3*. Curve *2* represents the discharge of capacitor *C2*. The time of *C2* discharge is so arranged that when curves *1* and *2* are added together, they produce a resultant linear curve as indicated by curve *3* in the drawing.

The circuit of Fig. 9-100A illustrates another simple method of linearizing a sawtooth wave by the addition of a new time-constant network. Resistor *R2* and capacitor *C2* are the added components. Nothing else in the circuit is changed from the conventional arrangement of Fig. 9-99A. This new circuit operates upon the same basic principles as that just discussed, whereby the linear trace of the resulting sawtooth depends upon the charging of one capacitor and the discharging of another. The circuit operates as follows:

Let us assume a starting point with capacitors *C1* and *C2* fully charged up from the B supply. The input signal biases the tube above cutoff, thereby making it conduct. The moment that plate current begins to flow, *C1* and *C2* both start to discharge through the tube. Capacitor *C1* discharges very rapidly as compared to capacitor *C2*, because the high-valued resistor *R1* is included in the discharge path of *C2*. As a result of this, by the time the tube becomes nonconductive, capacitor *C2* has only discharged a small portion of its voltage, whereas capacitor *C1* is almost completely discharged. Once the tube becomes cut off, *C1* charges up again from the B supply through *R1* and *R2*. However, the voltage across *C2* is much greater than that across *C1*, so *C2* still continues to discharge, its discharging current piling up electrons on *C2* and further charging this

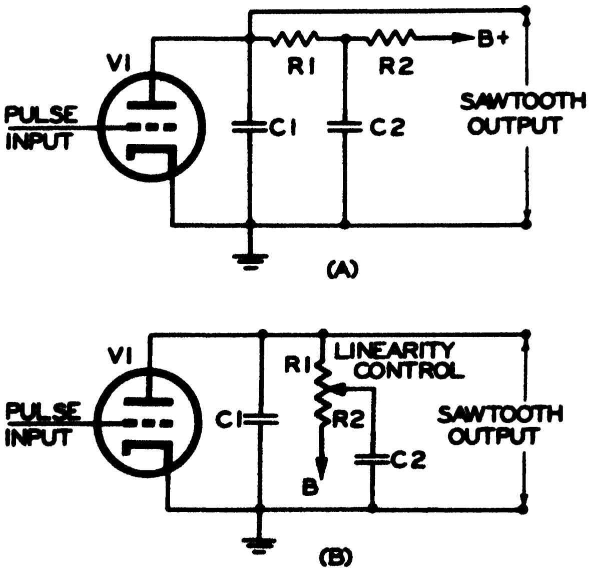

Fig. 9-100.—A simple circuit for improving the linearity of the sawtooth wave by means of an additional *R-C* network is shown in (A). Part (B) is a rearrangement of this circuit replacing resistors *R1* and *R2* by a potentiometer.

capacitor. A point will soon be reached where the voltage across *C1* and *C2* will be equal; *C2* will stop discharging and then begin charging through *R2* from the B supply. Thus both capacitors are once again in a state of charge and when the tube becomes conductive once more, the cycle of operation starts all over again.

With the output sawtooth signal taken from across capacitor *C1 only,* the reader may at first wonder how the operation of this circuit will linearize the sawtooth signal. The answer to this is easily explained if we carefully consider the two charging actions on *C1*. We said that *C1* charges up from the B supply and for a certain period of time also charges up as a result of the discharge of *C2* through *R1*. The charging of *C1* from the constant-voltage source is the normal rising exponential curve. The discharging voltage of *C2* is exponentially decreasing and part of this voltage is used to charge *C1*. Therefore, across *C1* we effectively have two charging voltages, that from the B supply having a convex shape and that from capacitor *C2* having a concave shape. The time constants of the circuit are so arranged that the superposition of these two curves produces a fairly linear sawtooth signal across *C1*.

The circuit of Fig. 9-100B is a rearrangement of Fig. 9-99(A). In this circuit, resistors *R1* and *R2* are replaced by a potentiometer; the arm represents the junction point of these two resistors, to which capacitor *C2* is connected. By the use of a potentiometer, it is easy to adjust the time constants of the circuit for the best possible linear sawtooth signal. The potentiometer in a circuit such as this is known as a "linearity" control.

"Bootstrap" Feedback Circuit

There are many different types of feedback circuits that are used to linearize a sawtooth sweep. One of the most common makes use of a cathode-follower circuit to correct the nonlinearity. Linearizing circuits utilizing cathode followers are found in television receivers as well as in oscilloscopes. A simplified drawing of such a circuit appears in Fig. 9-101. This circuit is often called a *bootstrap sweep*. *V1* represents the discharge tube which conducts only for a short period of time. Tube *V2* is the cathode follower. Although a sawtooth voltage exists across capacitor *C1*, the output sawtooth voltage is usually taken across the cathode resistor of tube *V2*, as indicated in the drawing. Before we analyze the operation of the circuit, let us study a few of the features about a cathode follower that are of interest to us.

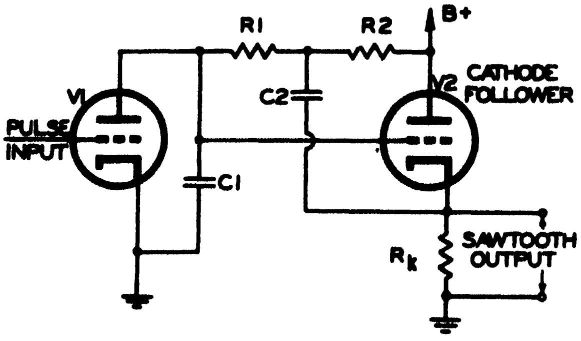

Fig. 9-101.—A feedback circuit to linearize the sawtooth sweep, often called a *bootstrap sweep*.

The load in the cathode circuit of a cathode-follower stage is unbypassed, with the output signal taken across this load. In the drawing of Fig. 9-101, resistor R_k represents the cathode load. The grid functions in the normal manner in that it receives the input signal in the usual manner, but the plate does not have any load resistor. The gain of a cathode-follower stage is always less than unity. The primary reason for this is that the signal voltage developed across the unbypassed cathode resistor is of opposite polarity and effectively in series with the signal on the grid, and degeneration occurs. Since all the signal voltage flows through R_k, the maximum amount of degeneration will occur. However, there is little or no frequency discrimination and whatever attenuation does occur is the same for all frequencies within the pass-band of the circuit. Thus, the relative amplitude at the different frequencies remains constant even though degeneration does occur and the over-all frequency response in the output is unaffected.

The gain of a cathode-follower stage is expressed as

$$A = \frac{R_k}{\frac{1}{g_m} + R_k} \qquad (9\text{-}26)$$

where g_m is the transconductance of the tube and R_k is the cathode resistor. If the g_m of the tube is high enough, or if R_k is of a high value, such that the expression $1/g_m$ is much smaller than R_k, the gain of the stage approaches unity. However, since $1/g_m$ cannot become zero due to the finite value of g_m, then the over-all gain of the stage can never be greater than unity.

The output impedance of a cathode-follower stage is *not* equal to the cathode resistor R_k alone. Due to the degeneration in the system, it can be shown that the equivalent output impedance, Z_L, is actually

$$Z_L = \frac{1}{g_m + \frac{1}{R_k}} \qquad (9\text{-}27)$$

From this latter equation, we can see that the transconductance of the tube is an important factor in determining the output impedance. With proper choice of the tube and the value of R_k, the effective load impedance of a cathode-follower stage can be designed for matching purposes. The effective load is always smaller than the chosen value of R_k. Thus R_k may be large enough to give the correct bias for operation of the tube but the effective load may be small if the g_m of the tube is high, thereby providing a low-impedance output.

It can also be seen that if the cathode resistance is made quite high, as well as the g_m of the tube being high, the quantity $1/R_k$ will be very small and may be considered negligible compared to the g_m of the tube. In this case, the effective load impedance becomes approximately equal to the reciprocal of the transconductance of the tube

$$Z_L \approx \frac{1}{g_m} . \tag{9-28}$$

Thus, if the cathode resistor is made high enough, the effective load will be independent of it and solely dependent upon the g_m of the tube. By making the effective load independent of the cathode resistor R_k, a more constant impedance output from the cathode follower can be obtained.

We do know that the signal potential across the unbypassed cathode resistor follows whatever varying potential is applied to the grid circuit. Thus in the "bootstrap" circuit of Fig. 9-101, the grid is connected directly to the high side of charging capacitor *C1* and the cathode potential follows this charging voltage. You will notice in the drawing that the cathode of *V2* is tied to the junction of resistors *R1* and *R2* through capacitor *C2*, so that signal voltage from this cathode circuit is fed back to charging resistor *R1*. If the gain of the cathode-follower stage is so arranged that it is very close to unity, the voltage appearing at the cathode of *V2* will very nearly equal that across *C1* during the periods of *V1* cutoff.

Capacitor *C2* is charged up from the B supply and, if its value of capacitance is high enough, the potential across it will remain fairly constant. Therefore, the complete voltage across R_k is fed back to the junction of *R1* and *R2* with a minimum amount of loss. This means that the voltage across charging resistor *R1* will be substantially constant and this will result in a constant charging current for *C1* flowing through *R1*. This results in a linear rising characteristic of the sawtooth sweep.

The voltage across *C2* does not remain entirely constant because, during the sweep, *C2* discharges slightly. This feature, coupled with the fact that the gain of the cathode follower is actually less than unity, causes the voltage across *R1* to decrease toward the end of the trace of the sawtooth signal. This means that the resulting sawtooth signal is not perfectly linear, but has some curvature at the end of its trace.

The resulting sawtooth signal nevertheless is quite linear as compared to the case where a cathode follower is not employed. It can be shown that the same linear sawtooth wave would result if capacitor *C1* were being charged from a high potential source several hundred times greater than the actual applied voltage.[19] It can also be shown that the charging resistance of capacitor *C1* is effectively increased by a factor $1/(1\text{-}A)$ times the actual charging resistance, where A is the gain of the stage.[20] It is interesting to note that since A is always less than unity, the effective charging resistance will always be greater than the actual resistance. If the gain of the stage approaches unity, the denominator in the expression $1/(1\text{-}A)$ approaches zero and the effective resistance will approach infinity.

By taking the sawtooth output voltage across the cathode resistor R_k in Fig. 9-101 instead of across *C1*, the cathode-follower stage can act as a buffer amplifier and also as an impedance-matching stage to the circuit following it.

Linearization By Nonlinear Amplifier

Another common method of linearizing the sawtooth sweep is to use a nonlinear amplifier. Such methods of linearity correction are not immediately noticeable at first because the correction amplifier is usually one of the deflection amplifiers of the system. By nonlinear amplifier we mean one that has curved dynamic transfer characteristic ($i_b - e_g$ curve). Such amplifiers are often referred to as logarithmic amplifiers because the shape of their transfer characteristic is similar to that of a logarithmic curve. Amplifiers that have such characteristic shapes generally utilize variable-mu, or remote-cutoff, tubes.

By feeding the exponential rising sawtooth voltage to the grid of the amplifier, and adjusting the amplifier's transfer characteristic so that its curvature is approximately opposite to that of the input exponential curve, the resulting sawtooth output signal will be fairly linear.

The drawing of Fig. 9-102 will make this clearer. Curve *A* represents a portion of the transfer characteristic of a variable-mu amplifier tube, and curve *B* the nonlinear sawtooth signal input to the amplifier. The bias on the amplifier is so adjusted that the input sawtooth signal operates over that portion of curve *A*

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[19] Puckle, *op. cit.*, pp. 94-96.

[20] M.I.T. Radar School Staff, "Principles of Radar," 2nd ed., pp. 3-18 to 3-20, McGraw-Hill Book Co., Inc., New York, N. Y., 1946.

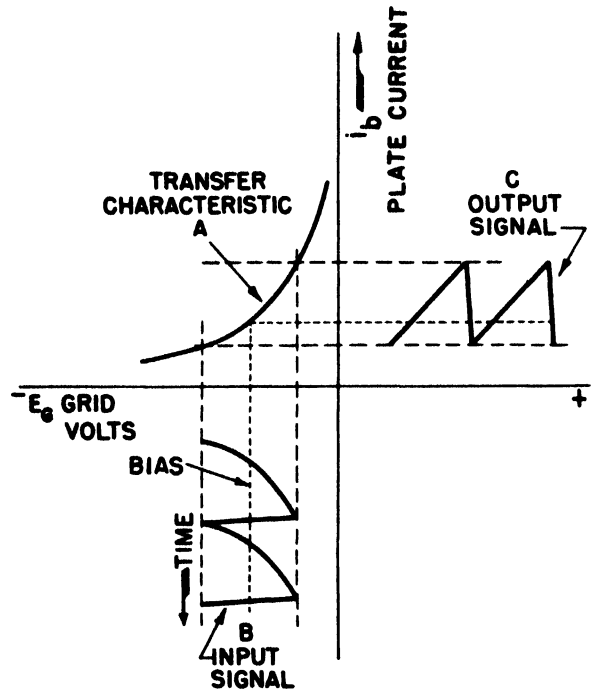


Fig. 9-102.—Graphical analysis of the linearization of the sawtooth sweep by use of a nonlinear amplifier.

that helps produce a linear rising plate-current output. The result of this operation is indicated by the linear sawtooth at *C* in the drawing.

A typical circuit utilizing such an amplifier appears in Fig. 9-103. In this drawing, *V1* represents the tube through which *C1* discharges to produce the fast retrace or flyback time of the sawtooth. Capacitor *C1* is charged from the B supply through resistor *R1*, and the resultant signal across *C1* is seen to be a sawtooth wave with an exponential trace which characterizes the nonlinearity of the sweep. This nonlinear sawtooth sweep is resistance-capacitance coupled by *R2-C2* to the grid of tube *V2*, which functions as the nonlinear amplifier. The plate voltage on this tube is adjusted so that the best possible transfer characteristic will be produced to correct the nonlinearity of the incoming signal.

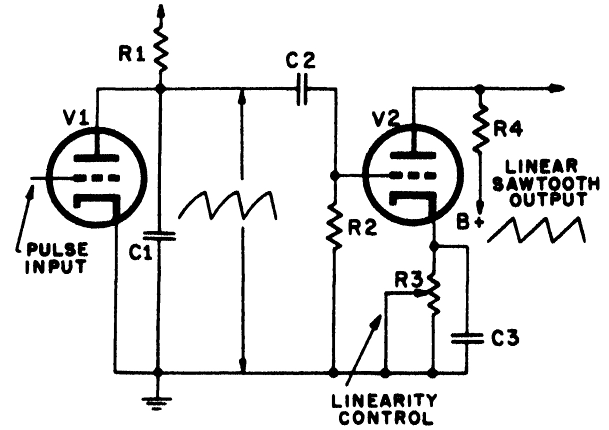


Fig. 9-103.—A typical circuit using a nonlinear amplifier for improving the linearity of a sawtooth sweep.

In many circuits, cathode bias is used on these amplifiers. By employing potentiometer *R3* in the cathode circuit, the bias on the tube can be adjusted so that the incoming signal can be made to operate over the best possible part of the transfer characteristic. In this manner, the linearity of the output signal can be somewhat regulated, and thus *R3* serves as a linearity control. The linear sawtooth signal output is taken from the plate circuit of the tube. Component *R4* represents the plate load for this tube.

CHAPTER 10

THE BASIC OSCILLOSCOPE AND ITS MODIFICATIONS

There are in use a great number of different types and models of oscilloscopes. All are alike in some respects, but each has something in it or is labeled in a manner which makes it different from the others. Sometimes it is a technical detail and other times it is the selection of the names which identify the controls. Nevertheless there must be a common meeting ground for all of these instruments. This is the basic oscilloscope, a device containing the minimum number of essential sections which, when combined, form a unit of practical usefulness. The expansion of these sections and the addition of other special circuits result in one or more of the different varieties of more elaborate equipment.

We will assume no familiarity with the basic oscilloscope on the part of the reader. The preceding chapters have established some background, but references to the device as a whole, as for example in Chapter 1, were very general in nature. It disclosed very little of what constituted the finished device. Chapters 2 through 9 treated the theory underlying the various actions which occur within the unit, but the correlation of these actions has yet to be given. It shall be done in the present chapter.

THE BASIC OSCILLOSCOPE

The basic oscilloscope is not a commercial instrument. It is an ideal device containing the various features common, and therefore necessary, to all oscilloscopes. The basic oscilloscope is a working device. This it must be if we are to show how the sections perform and what one can expect to see on the screen under different circumstances. But the absence of standardization of oscilloscope nomenclature makes it impossible to give names to the different parts of the basic instrument which will immediately correlate it with any real instrument.

The problem is not in the circuits within the oscilloscope cabinet. These are very much alike, in general, in all oscilloscopes, and, like radio receivers, differ only in minor details. The addition of circuits to expand the capabilities of the instrument presents no obstacles to the analysis, either. The problem that faces the reader is the association of the names of the elements of the basic device with their counterparts in the real instrument which one is studying or operating. We shall make every effort to aid in such correlation.

In order to most clearly present the fundamental details of the commercial oscilloscope, we shall treat the basic version as if it were divisible into a number of major and minor sections. This treatment will aid in associating the elements being discussed here with the elements in the device which the reader is studying or using. Possible confusion due to the different names assigned to these sections in the commercial product is, therefore, minimized. All sections of the device are not always in use, although some are common to all applications. Therefore, the combinations of sections must be identified in order to distinguish one arrangement from another in a particular application of the instrument. This is particularly pertinent in the choice of the switches which arrange the required combination of active sections, and also the variable controls which are associated with these sections and play a very important role in the proper operation of the unit.

Sections of the Basic Oscilloscope

The basic oscilloscope can be divided into six major sections and three minor ones. The distinction between the two groups is based on the complexity of the circuits associated with each, rather than on the contribution made by each to the end result. It is a breakdown of convenience rather than an attempt to indicate the relative importance of the various circuits. By a major section we mean an assembly of tubes and circuit components which perform a broad function; by a minor section we mean as few as one or two circuit components, as for example, a potentiometer, or perhaps a tube and a few related components. The breakdown is obviously arbitrary and solely for the purpose of enabling a certain form of presentation. Once the functions of the sections are understood, the division of duties as we describe them can be forgotten on an individual basis and the instrument can be treated as a whole. The selection of the proper controls and switches associated with these sections will become automatic, for these rather than the sections themselves enable the device as a whole to perform its duties. These are the switches and the controls that correlate the sections. But we must deal with the larger

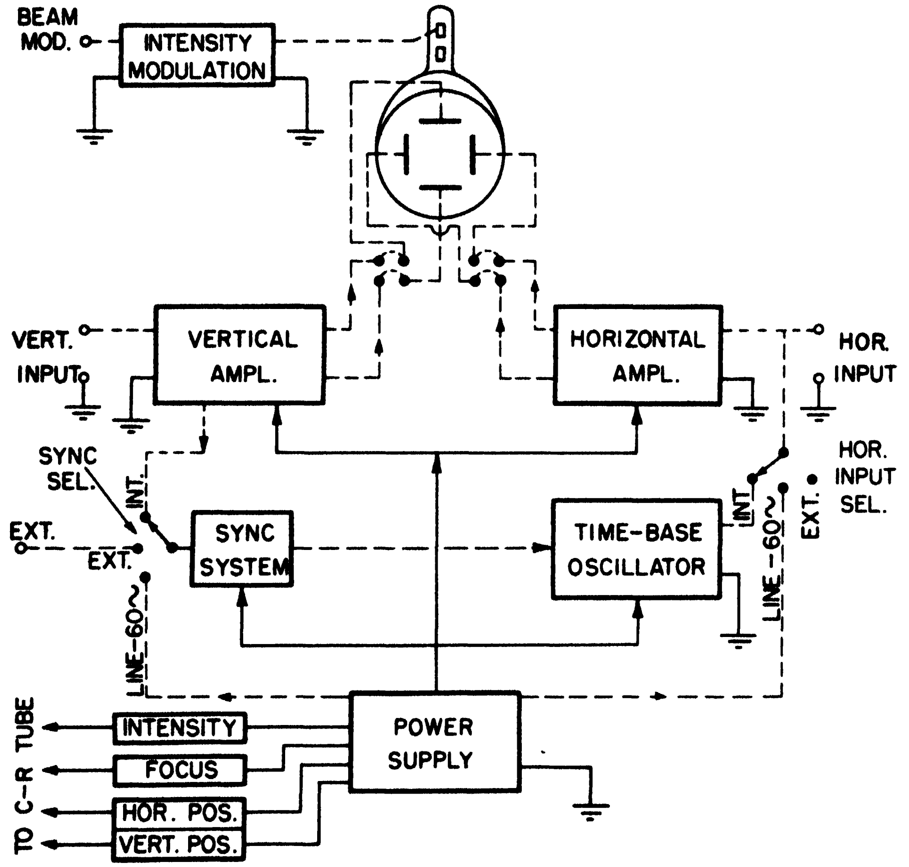


Fig. 10-1. — Block diagram of basic oscilloscope.

segments first for they are all-inclusive; then we can subdivide them into their subsections.

When listed in the order which permits their most logical discussion, these nine sections appear as follows:

1. The cathode-ray tube
2. The beam intensity and focusing system
3. The beam-positioning system
4. The vertical-deflection amplifier
5. The horizontal-deflection amplifier
6. The time-base system
7. The synchronizing system
8. The intensity-modulation system
9. The power-supply system.

All of these, with the exception of the cathode-ray tube, are functional names; they indicate, in the most general terms, what each system does as a part of the over-all action. Physically, all are located within the confines of the oscilloscope housing, and are accessible only by going inside the cabinet. Electrically, on the other hand, they are accessible, in most instances, from the face of the cabinet by means of switches and other controls which are mounted there.

When illustrated in block diagram form, these nine sections of the basic oscilloscope are as they appear in Fig. 10-1. Each block has only one significance at this time, namely, the basic function indicated. The full meaning of these functions may not be clear yet; they will become so as the chapter unfolds. Because of their association with signal circuits and with operating voltage circuits, the blocks are tied to each other with different kinds of lines. The signal circuits are shown by dotted lines and the power-supply circuits are shown by solid lines. The arrowheads indicate the direction of application of the signals and the operating voltages. The absence of continuity between the power-supply section and the cathode-ray tube via the intensity, focus, and beam-positioning controls is occasioned by the desire to simplify the drawing. The single-line signal paths are possible because of common grounds at all points except at the output of the vertical and horizontal amplifiers. This is in contrast with the early versions of cathode-ray oscilloscopes, the devices produced during the early 1930's. In those instruments, one side of the amplifier output circuit was grounded. The practice has not been entirely discontinued and some instruments, few though they may be, still employ such circuits. Every device which has a single-ended output in the vertical or horizontal amplifier has one of the output terminals at ground potential.

Before embarking upon the discussion of the functions of the sections, let us call attention to certain very significant details concerning the contents of Fig. 10-1. Everything of interest inside the device is not evident in this drawing, but two items deserve attention. These are the two switching arrangements. Because it would be premature at the moment to deal with names or labels for these elements, let us identify them as the switch adjacent to the block labeled SYNC SYSTEM and the switch adjacent to the block labeled HORIZONTAL AMPL.

Each of these switches has three positions. In the former, the position INT connects SYNC SYSTEM (whatever it may be) to the VERTICAL AMPL; the position EXT seems to go nowhere (temporarily only), and the position 60~ seems to connect SYNC SYSTEM to the POWER SUPPLY. This latter connection is fairly obvious, being the 60 cps somewhere inside the power-supply unit.

The switch near TIME-BASE OSCILLATOR connects HORIZONTAL AMPL to TIME-BASE OSCILLATOR when set to INT position, to POWER SUPPLY when set to 60~SINE and apparently nowhere when in position EXT.

We have a definite reason for focusing attention upon these two switches. Obviously they involve a number of sections, five of them, specifically: the vertical amplifier, the synchronizing system, the power supply, the horizontal amplifier, and the time-base oscillator. The switches permit various combinations of these sections. Each of these combinations has a definite purpose and the switches are set at specific positions in accordance with the intended duties of the oscilloscope, thereby setting up the various combinations of sections necessary for the accomplishment of the desired results. The reason for pin-pointing the switches, which we have already mentioned, but an additional and equally important reason, is the fact that manufacturers of oscilloscopes have elected to assign different labels to these switches, and any attempt at the moment to correlate them with commercial equipment must be made in terms of what they do, rather than the names by which they are known. In fact, this situation is so aggravated that a separate illustration will be referred to later as the means of correlating the different nomenclatures with the sections they signify. But let us first deal with the broad details of the sections.

The Cathode-Ray Tube

The cathode-ray tube is the display element. By this time, we know that the electrical equivalent of the phenomenon being studied appears on the screen of the tube and additional details of this kind would be superfluous. Much remains to be said, however, about the trace which appears on the screen. The actions of the different sections and controls within these sections have a profound influence on the character of the display.

In this connection, Fig. 10-1 discloses that the tube (the deflection systems) receives signal voltages from two sections of the oscilloscope, the vertical and the horizontal amplifiers. These signals are carried to the deflection plates through the usual conducting wires, but *sometimes* they pass through jumpers which link connecting-point terminations of the deflection plates and the output systems of the two amplifiers. These are shown by the two pairs of small circles located in the paths of the amplifier output connections to the deflection plates. These eight connecting points usually are found mounted on a terminal block, or perhaps four on each of two blocks, which are accessible at the rear or the sides of the oscilloscope cabinet. Sometimes they are located inside the cabinet.

The remaining parts of the cathode-ray tube, that is, the heater, control grid, and other elements of the electron gun, receive their operating voltages from the power supply. Among these voltages are those required for static positioning of the electron beam, and naturally, the voltages which govern the intensity and the focus of the beam. These are the four small blocks which are tied to the power supply and bear arrows indicating termination on the appropriate points of the cathode-ray tube. Inasmuch as the contributions made to intensity of the trace and the focusing of the beam by the different elements of the electron gun have been described in a previous chapter, we shall refrain from further comments concerning the behavior of the electrodes. The visible effect on the trace, however, will be discussed shortly.

The screen of the cathode-ray tube is visible from the front of the cabinet, as was shown in numerous illustrations in Chapter 1. In connection with the screen, we might mention that sometimes a retractable light shield surrounds the tube screen and can be pulled out of or retracted into the cabinet. Its function is to shield the screen of the tube against incident light, thus permitting greatest visibility of whatever trace appears on the screen. Also, a circular transparent plastic or celluloid sheet bearing a number of crossed lines, which is attachable to the face of the tube by means of a number of spring clips, may be in front of the tube face. Since we shall show examples of traces as they appear on the screen of the tube, this calibration scale is assumed removed; otherwise the traces illustrated in this chapter will not be the same as those seen on the screen. When viewed through the crossed-line sheet, the traces appear to be broken, although this does not impair the ultimate utility of the calibration chart. So much for the details concerning the cathode-ray tube. The appearance of the trace will be discussed separately.

The Beam Intensity and Focusing System

It is advisable to discuss the intensity and the focusing systems together, even though the design of the modern-day cathode-ray tube is such that these two actions are somewhat independent of each other. But even under the best of conditions, a change in one will somewhat affect the state of the other.

When intensity is referred to, it means two things, although in the end it still means the visibility of the trace when seen on the screen of the tube. Inside the tube, it means the electron density of the electron beam. The greater this density, the greater will be the brilliance of the trace when seen on the screen. Practically speaking, it is the latter which deserves our attention, and because it also is influenced by another control, we must of necessity speak about the result of the focusing action at the same time.

Inasmuch as we have shown in Chapter 8 that the trace is formed by a single dot of light which develops at the point of impact of the electron beam on the screen, we can explain the actions of variations in intensity and focusing in terms of the appearance of this luminous dot. In the main, its luminosity or brilliance is a function of the intensity adjustment (control), as is its size, but the correctly developed dot is the one which is smallest in diameter, being circular in shape, and is most brilliant. This cannot be achieved by means of the adjustment of intensity only; it also requires correct focusing or adjustment of the focus control.

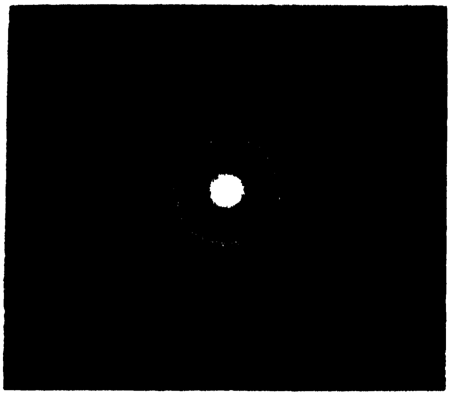


Fig. 10-3.—Spot with halation.

An example of a correctly focused dot is shown in Fig. 10-2. It is impossible to illustrate intensity, but it can be described by saying that the *minimum* intensity consistent with adequate visibility under the operating conditions and proper focus results in the correctly proportioned spot dimension as well as in the spot of correct brilliance. This comes about as the consequence of the fact that increasing the intensity will increase the spot dimension with or without correct focusing. Another visible condition also accompanies the variation in intensity, namely, the development of one or more secondary rings of light around the primary spot. These are known as halos, and their appearance is shown in Fig. 10-3.

The use of the minimum intensity and correct focusing cannot be set down as a rigid rule under all conditions of use; incident light on the face of the cathode-ray tube will reduce the visibility and demand that greater intensity be used. This may result in a halo even when the electron beam is properly focused, but if the screen is being viewed under what might be considered special conditions, that is, from a relatively great distance under brightly lighted conditions, the halo will tend to disappear from view. At least it will not be bothersome to the eye.

Such operation is not the best as far as tube screen life goes, but circumstances alter conditions. It is important, however, even then, not to use greater intensity than is necessary to achieve the desired visibility; moreover, the better the degree of focus, the less will be the required electron density in the beam in order to achieve greatest brilliance.

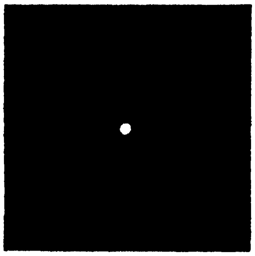


Fig. 10-2.—Properly focused spot.

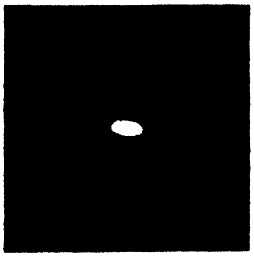


Fig. 10-4.—Unfocused elliptical spot.

Concerning the normal focusing action, the degree of focusing has a bearing on the size of the dot of light, and on its roundness as well. As a matter of fact, an unfocused spot will more than likely be elliptical in shape, with its axis either vertical or horizontal, as shown in Fig. 10-4. It will have the outline of an egg held upright or lying on its side. Neither one is acceptable, and they must be corrected. With this departure from roundness there also takes place an expansion in its dimensions. Fig. 10-5 shows an unfocused spot which is circular in shape.

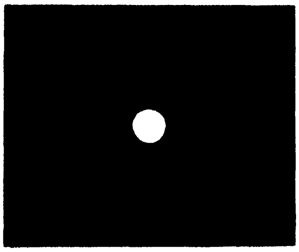


Fig. 10-5.—Unfocused circular spot.

Summarizing the functions of the beam intensity and focusing systems, we note that they affect both spot dimension and brilliance. Applying this to the practical application of the tube, we might well comment that even under correct conditions, all tubes are not designed to give equal-sized traces of the beam at the point of impact. While it is true that even in the case of the largest spot, it is a very small fraction of an inch in diameter, as described in Chapter 2, all cathode-ray tubes cannot be adjusted to give the same spot size.

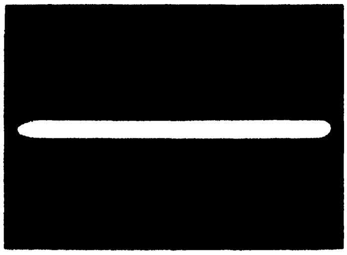


Fig. 10-6.—Horizontal trace with intensity too great.

Another very important point to bear in mind is that the static (undeflected) spot brilliance will not necessarily be the correct one when the electron beam is deflected and the spot moves across the screen. The size and the brilliance of the spot are related to the area of the time during which the screen material is excited by the electron beam. If the beam is in motion, the period of screen excitation is reduced at every point where the beam strikes the screen, being least at that point in the cycle of the phenomenon where the rate of change is greatest. In general, the brilliance of the trace will be materially reduced. This means that the intensity setting must be increased, and each increase in intensity will demand a readjustment of the focus. Excessive intensity under such conditions will result in a traveling halo and will tend to make examination of the trace more difficult. Purely as a means of conveying an impression, Fig. 10-6 illustrates the appearance of a straight-line trace (horizontal deflection) when the beam density is too great, and Fig. 10-7 shows such a trace when the spot dimension is correct. The line in the former is too broad and would impair analysis of a quantity which involved numerous variations in amplitude during each cycle.

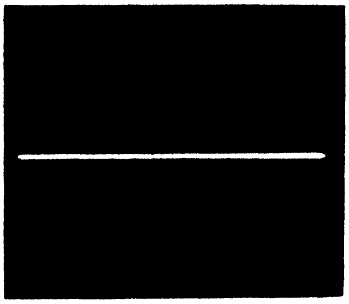


Fig. 10-7. — Horizontal trace with proper intensity.

Another item of a practical nature relates to the life of the screen when an undeflected beam of high density (high brilliance) is allowed to remain stationary at one point on the screen. It can very rapidly burn the screen. This is an irreparable damage, and while it does not impair the utility of the other parts of the screen, it might tend to cause annoyance when examining a trace. So, if the beam is to remain stationary on the screen for a prolonged period of time, the intensity must be reduced. It is an excellent precautionary measure. The beam need not be extinguished completely; if it is just visible, it will not damage the screen coating. High intensity is undesirable also when the beam is in motion and is retracing the same path over and over again, especially when the tube is displaying a low-frequency action. Again we stress that the intensity should be the lowest consistent with visibility. High intensity when the beam is traversing the same path at a rapid rate will not damage the screen, but the lower the intensity, the less will be the possibility of damage under any circumstances.

The Beam-Positioning System

The duties of these systems are expressed by the heading of this paragraph, although several interesting details may be given. Reference to Fig. 10-1 discloses that the two systems appear in the basic oscilloscope, and we might well add here that they are common to all such devices regardless of their category. One affords control of the static position of the beam along

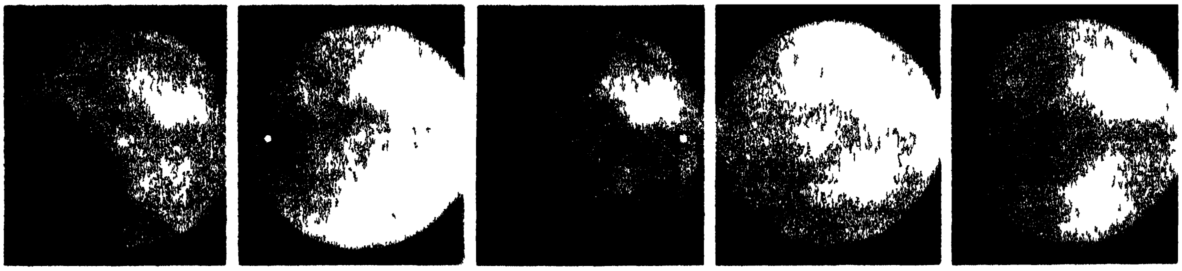


Fig. 10-8.—Various undeflected spot positions

the vertical axis of the screen, and the other gives control of the beam location along the horizontal axis of the screen. Reference to the "static position" should not be construed to mean that the controls are ineffective when the beam is in motion, for, as a matter of fact, they not only are very useful then, but the frequency of their use in that manner is as great as, if not greater than, when the beam is undeflected.

Since the normal process of deflection of the electron beam has been established, it is quite in order to say that the beam-positioning systems are means whereby the two plates of each deflection system are subjected to a fixed d-c voltage which establishes a deflection field between each pair of plates. These fields are the mean values around which the normal a-c signal-deflection fields vary, much as a signal applied to a biased grid will cause that grid to vary in potential around the steady bias value. When the signal-deflection fields are absent, the steady d-c fields locate the electron beam relative to the screen surface, according to the intensity of the positioning field. This point corresponds to the zero-voltage value of the signal-deflection field, so that by varying the positioning voltage field, the trace as a whole can be shifted to any point on the screen which suits the purpose of the viewer.

By means of a positioning voltage applied to the vertical plates, the point of impact of the electron beam on the screen may be moved vertically across the screen surface, and by means of the positioning voltage applied to the horizontal-deflection plates, the beam may be moved horizontally along the screen. While these words describe the action in general terms, it might be well to remark that this motion may be limited by other undesirable conditions within the oscilloscope. This should not be taken to indicate a deficiency in the beam-positioning circuits, but rather as the explanation of the performance specification which states the maximum linear movement of the beam possible by means of the positioning systems. Certain circuit designs introduce complications when the beam-positioning controls are advanced beyond certain reasonable limits, although the amount of beam displacement normally allowed is more than ample for all conventional uses of the tube. Strangely enough, these effects are encountered more frequently in the elaborate versions of the oscilloscope than they are in the basic units, but only when the stated limits of beam positioning are exceeded.

Before concluding the subject of beam-positioning systems, we are anxious to emphasize the fact that when beam positioning is used when the beam is undeflected, usually it is restricted to the location of the spot at the geometric center of the tube. Seldom is the location of a trace of an action positioned on the screen by moving the undeflected spot; the normal procedure is to position the complete trace. In that way, the number of readjustments necessary for the display of the complete wave is kept to a minimum. Although the meaning of different static beam positions on the screen of a cathode-ray tube has been intimated on a number of past occasions, Fig. 10-8 shows several views of the cathode-ray-tube screen with the undeflected post variously positioned. The correct one under conditions of static deflection, that is, in the absence of deflection, is at the center of the screen.

The Vertical-Deflection Amplifier

The vertical-deflection amplifier, also known as the Y-axis amplifier, is the path by which the signal being studied generally reaches the vertical-deflection plates. This qualification is necessary because under certain conditions, the vertical amplifier is not used; however, for the present, we shall imagine that it is the path for the signal.

Essentially, it is an amplifier or group of amplifiers, possessing specific capabilities and beset by the usual limitations of amplifiers. It is the means of raising or lowering the signal level to that required by the deflection system for a trace of useful proportions. Physically speaking, the entire amplifier is located inside the cabinet, but its related controls are mounted on the panel of the instrument. These are associated with the input system of the amplifier, whereas the output connections either terminate at a terminal board which also carries the connections for the two vertical-deflection plates of the cathode-ray tube, where they are linked by means of jumpers (as shown in Fig. 10-1), or they are connected directly to the deflection plates inside the oscilloscope unit.

Signal input to the vertical amplifier is accomplished in two ways, either through the use of binding posts or by means of a connector socket which will accommodate the male plug at the end of an input cable which allows for a comparatively remote location of the signal source. The latter arrangement, however, is not always a part of the equipment, whereas binding posts are standard in almost all cases, and are the means used in the basic oscilloscope. We might mention that in a few isolated cases, binding-post connectors are replaced by pin-jacks.

The amplitude of the signal which issues from the vertical amplifier is controlled by the *vertical-amplifier attenuator* and for the *vertical gain control* which is in most cases located at the input of the amplifier and determines the magnitude of signal admitted *into* the amplifier proper. This portion of the amplifier is represented on the front panel of the oscilloscope by either one or two controls, depending on the design of the input system of the vertical amplifier, or more specifically depending upon the design of the attenuator system. In general, this element appears as a single potentiometer, a step voltage divider, operating in conjunction with a continuously variable potentiometer, but in a few examples of the basic oscilloscope, it consists of two units. Whichever it may be, the fact remains that it is the means for controlling the signal input to the amplifier, and in that way the signal output from the amplifier, the latter being a fixed gain system.

The input signal attenuator and the vertical amplifier are connected so that the attenuator is not available for use without the amplifier; neither is the amplifier available for use without the attenuator, although, as would be expected, the latter can be adjusted by proper setting of the controls so that its presence in the circuits has negligible effect. In other words, it does not attenuate the signal; this is like any other attenuator system used with an amplifier.

In the light of the normal function of an amplifier, it might be well to comment that the vertical amplifier in the oscilloscope performs a dual function, or rather the attenuator in the amplifier performs these functions. Normally an amplifier is required to raise the signal level, but sometimes the conditions of operation are such that the signal to be fed to the vertical-deflection plates for display exceeds the required value; the vertical-amplifier input control is then needed for reducing the signal level to the value necessary for the display of a trace of suitable proportions. Under the circumstances, the vertical amplifier always should be in the circuit (if its performance characteristics allow it), even if its only purpose is to *reduce* rather than increase the signal-voltage level.

As a final comment for the present, we might remark that the vertical-amplifier attenuator is a much more important device than may appear from the casual treatment we have given it. In fact, that part of the vertical amplifier is so important and so far-reaching in its effects that it receives very special attention later in this chapter. Normally in most oscilloscopes of the basic variety, the vertical-amplifier attenuator is a simple potentiometer type of voltage divider. It is entirely adequate for the investigation of sine-wave phenomena and even some nonsinusoidal waveforms, although the latter demands more serious consideration than the former. As a control of sine-wave signal amplitude, the device is simply a signal-level control and its frequency characteristics are of little importance. In other kinds of work, however, its frequency effects may be serious, sufficient to impair the usefulness of the display. Because of this, the units designed for more serious research work in the realm of high frequencies and pulse techniques make use of a more elaborate attenuator network. This is shown and discussed elsewhere in this chapter.

The vertical amplifier in an oscilloscope is a complete system and must be used as a complete unit. Facilities for varying the number of tubes used for amplification, or the operating voltages, generally do not exist. All the circuits are fixed and remain that way during use. At the moment we know of but one exception to this statement and this only in a very elaborate version of the oscilloscope. Whether or not the practice will become more prevalent in the future, no one knows.

In view of the fixed gain nature of this amplifier and the fact that the input signal is varied in level, one can readily appreciate the existence of a minimum input signal level in order to develop an output signal of suitable magnitude. This would correspond to the input voltage with zero attenuation. As a rule, this is not difficult to secure, but the occasion may arise when it is not available, in which event, either an additional amplifier is added between the signal source and the input of the oscilloscope amplifier, or the oscilloscope cannot be used for the purpose of observing that signal.

While on the subject of signal voltages at the input of the vertical amplifier, it might be well to state that each such amplifier bears a rating of the *maximum* voltage which may be applied across the input terminals. This usually refers to the d-c voltage between the points to which the input of the amplifier is connected. In that way, breakdown of the isolating capacitor used at the input of a-c d-c amplifiers is avoided, and d.c. is kept from the input tube. In the case of d-c or zero-frequency amplifiers, a similar rating exists, but this time in connection with the ability of the attenuator element to withstand the full d-c voltage which is applied across it.

Amplifiers which feed the vertical-deflection systems in oscilloscopes or in other devices which utilize cathode-ray tubes differ in many ways. One of these differences, namely, the frequency response or bandwidth, was discussed in Chapter 1. As far as the basic oscilloscope is concerned, the pass-band of the vertical amplifier is of little consequence. It has no bearing on *how* the device is employed, but it does influence the nature of the work which can be done with it, which is an entirely different matter.

As far as gain in the system is concerned, a variety of figures exist, although a fair approach to standardization is to be found among oscilloscopes intended for use in the same field, or instruments which are offered within any one price range. This is natural, inasmuch as similar applications dictate similar requirements; from the economic viewpoint, the same factors of cost and profit apply in virtually all manufacturing localities. In general, however, equipment capabilities increase with sales price if for no other reason than that buyers expect more for more money. As a factor determining cost, frequency bandwidth is very important; the greater the bandwidth of an amplifier, generally the higher is the cost of the instrument. The reason for this will become evident when we describe the means of obtaining wide bandwidths.

Another point of difference is the number of stages of amplification. This influences the frequency range of the unit, its over-all gain, and also the *phase* of the signal at the output relative to the signal at the input. Functionally, the latter condition is not a limiting factor in any way, because its only effect is to determine the direction of movement of the spot on the screen relative to the instantaneous polarity of the signal at the input of the amplifier. The only reason for mentioning this fact is that an identical signal fed into two oscilloscopes of different make may cause the spot to move in opposite directions.

The Horizontal-Deflection Amplifier

When considered alone, the horizontal-deflection amplifier or horizontal amplifier as it is more frequently known, or the X amplifier (which is still another name for it), should be considered as simply another amplifier. It derives its name from its association with the deflection system to which it normally channels signals applied to its input. Its output terminals connect directly to the horizontal-deflection plates, or to a terminal board where a pair of jumpers link the amplifier output with the two deflection plates.

Functionally speaking, it is the channel whereby the signal destined for horizontal deflection is either raised or decreased in level as required for the development of a trace of reasonable proportions on the screen of the tube. In this respect, it serves exactly the same purpose as the vertical amplifier. There are other points of similarity between the two, as well as points of difference.

The treatment of the signal applied to the input of the horizontal amplifier is very much like that found in the vertical-deflection channel. The signal may be applied via two binding posts, which are sometimes replaced by pin-jacks, or it may be fed to the amplifier via internal wiring controlled by a switching arrangement. Either way, the signal is under the control of the input attenuator and this element may be a single potentiometer, or it may be a two-section system consisting of a step attenuator and a continuously variable resistor. Whichever it is, the fact remains that the amplifier proper is a fixed gain system and the attenuator determines how much signal enters the amplifier, and, therefore, the magnitude of the signal which leaves the amplifier and is applied to the horizontal-deflection plates.

Unlike the vertical amplifier, the input of the horizontal amplifier may be tied in electrically with other sections of the oscilloscope, as indicated in Fig. 10-1 and as will be shown shortly. However, it also is able to receive a signal from a source external to the oscilloscope, this external signal being applied via the binding posts or the pin-jacks. Obviously, every signal amplified by the horizontal amplifier need not be the time-base signal. This is evident in Fig. 10-1 and also from the remarks in the last two paragraphs, but it can bear repetition because of the very close association between the horizontal deflection and the time-base voltage in oscillography.

The electrical characteristics which have been described in connection with the vertical amplifier find their parallel in the horizontal amplifier, for, after all, it still is an amplifier, even if its function in the oscilloscope differs from that of the vertical channel. Its number of stages is fixed so that its gain is fixed. In this respect, we know of no exceptions. So, when used as an amplifier for an external signal, minimum signal levels exist at its input relative to a trace of proper dimensions on the screen of the tube. This time the deflection factor of the horizontal-deflection system is the determining influence.

Like the vertical amplifier, the horizontal amplifier has definite frequency bandwidth, and this bandwidth exerts a great influence on the suitability of the device for a signal. It so happens that in the case of the horizontal channel, its bandwidth requirements are not as severe as those for the vertical amplifier, but only in connection with the amplification of the sawtooth oscillator signal. If as wide a range of signal varieties are to be passed through the horizontal amplifier as through the vertical amplifier, then the same basic requirements apply to both. This is not the case in practice, however, except for certain operations, which are, therefore, limited in scope if the two amplifiers bear unlike frequency ratings. Because this is the general practice, and since the fundamental purpose of the horizontal amplifier is the amplification of the sawtooth signal from the time-base oscillator, the frequency bandwidth rating of the horizontal system usually is less than that of its vertical counterpart. This was brought out in the general discussion in Chapter 1 but is elaborated somewhat here.

The problem of bandwidth is a major one in oscilloscope amplifiers. It stems from the requirements of nonsinusoidal-wave amplification. This subject is treated in full detail later in this chapter, but for the present let it be said that all nonsinusoidal waves do not have the same harmonic composition. Therefore, the ability of an amplifier to accept and pass such a wave and yield a reasonable reproduction of the wave at the output is determined by both the fundamental frequency and the number of harmonics of that frequency which the amplifier will pass, and, moreover, the number of harmonics which are *considered necessary* for acceptable reproduction.

In a few instances, the frequency bandwidth of the horizontal amplifier equals that of the vertical amplifier, but, as a general rule, the horizontal amplifier offers coverage at a lower frequency, but at the high end falls far short of the vertical amplifier. In the wide-band oscilloscopes, the ratio between the upper-frequency limits of the two amplifiers may be as low as 5:1 or they may be as high as 20:1, depending on the individual design. In the general run of low-band and medium-band units, what we might call the basic versions, the ratio may be from 1:1 to perhaps 5:1, in all cases the larger quantity designating the vertical amplifier.

There is a definite reason for this, and while it may be somewhat premature at the moment, we might mention it here. Certainly it will be more understandable by the time the chapter is completed. The first reason is the frequency bandwidth of the time-base oscillator, and we might just as well have placed the time-base oscillator first in explaining the reason for the reduced bandwidth of the horizontal amplifier. To begin with, an ideal portrayal of a quantity, taking all things into account, requires about three cycles to appear on the screen. This means that the time-base oscillator must have an output voltage at a frequency about one-third that of the lowest frequency which may be passed through the vertical amplifier. Sometimes this is true and sometimes it is not, the latter being the case when two cycles of the quantity under investigation is the basis used in design. Whichever it is, it explains why the horizontal amplifier extends lower in its frequency limit than the vertical amplifier, and immediately establishes the relationship between the low-frequency requirement of the time-base or sweep oscillator and the lowest signal frequency acceptable to the vertical amplifier.

At the high end of the horizontal-amplifier pass-band, we find a different situation. To explain all of this at this time is unfair because it would impose an undue burden on the reader since the explanatory text is yet to come, but we might make a few brief remarks which will at least shed some light on the matter. The basis of design for most oscilloscopes is square-wave rather than sine-wave reproduction. Any oscilloscope which will give a good version of a square wave of a certain fundamental frequency will reproduce a wave of many time that frequency, usually twenty to perhaps fifty times the frequency, depending upon the individual design. At any rate, the time-base oscillator frequency for the display of a number of cycles of the square wave or some other nonsinusoidal wave is determined by the fundamental frequency of that wave and *not* by the highest harmonic components contained in that wave. Moreover, the time-base oscillator output is a sawtooth wave and a good reproduction can be obtained (as will be shown later) with only seven harmonics of the fundamental, so that if an oscilloscope is described as being capable of displaying a 10,000-cps square wave which would call for a 200,000-cps top frequency in the vertical-amplifier pass-band, the horizontal amplifier need pass only about (10,000/3) × 7, or 22,000 cps, in order to properly display that wave, since the time base then would be adjusted to 10,000/3, or 3,333 cps.

If we are to become more severe in our requirements and say that a basic oscilloscope is capable of displaying a 50,000-cps square wave by having a vertical-amplifier pass-band of 1,000,000 cps, the highest frequency to be passed by the horizontal amplifier is only (1,000,000/3) × 7, or 220,000 cps, if we show three cycles of the square wave.

More elaborate oscilloscopes are designed with the idea of attaining better reproduction of a square wave than would be obtained when twenty harmonics were passed by the vertical amplifier. This is why their vertical amplifiers have such a very wide pass-band. But this would not alter the situation we have described relative to the requirements of the horizontal amplifier, or the time-base oscillator. If their vertical amplifiers have a higher top frequency, to accept square waves of higher fundamental frequency, the horizontal-amplifier bandwidth increases proportionately, in order to accommodate the oscillator frequency, which has been increased to give the same three cycles of the wave on the screen.

Obviously, the lower the bottom frequency of the horizontal-amplifier pass-band, the better, provided that the vertical-amplifier pass-band has a sufficiently low frequency limit, which usually is the case. By the same token, the higher the upper-frequency limit of the horizontal amplifier relative to the highest sawtooth frequency available from the time-base oscillator, the better. In most general-purpose instruments, this ratio is about 3:1, and considering the general run of applications, it seems adequate, although personally we would like to see a higher figure, if only because of the usual rating of the time-base oscillator. A high upper frequency in the latter and a relatively low upper-frequency limit in the former are not compatible. Here are some figures which show the vertical- and horizontal-amplifier bandwidths relative to the frequency range of the time-base oscillator in the same instruments. These few cases were selected at random.

Vertical-Amplifier Bandwidth (cps)	Horizontal-Amplifier Bandwidth (cps)	Time-Base Frequency Range (cps)
15-100,000	15-100,000	8-30,000
2-200,000	1-100,000	0.5-50,000
10-1,000,000	5-250,000	10-30,000
5-11,000,000	2-500,000	5-100,000
2-100,000	2-100,000	3-30,000

All of the oscilloscopes tabulated above are not basic instruments, although four of the six are, but it is clearly evident that these devices are designed for certain purposes. Take, for example, the oscilloscope with a horizontal bandwidth of 5-250,000 cps and a time-base oscillator frequency range of 10-30,000 cps. If we are to show three cycles of a signal on the screen, the lowest frequency fed to the vertical amplifier is limited to 30 cps, but this is satisfactory to the market for which this device is intended, namely, television maintenance, for which work a bottom limit of 30 cps to the vertical system is adequate. Of course, if a single cycle is to be shown, then the bottom limit of the vertical signal frequency is 10 cps. When used alone, the vertical and horizontal amplifiers will accept frequencies of 10 and 5 cps, respectively.

It might also be of interest to comment that the rated frequency limits of the vertical and horizontal amplifiers are not the actual limits; each amplifier will accept sine waves of low frequency, as is described later in this chapter. Relative to the amplification of sine waves, the meanings of all of these pass-band limits and even the frequency range of the time-base oscillator differ, but we must hold the discussion of those meanings in abeyance until later in this chapter, when we have described certain operations which permit the examination of fewer cycles than are indicated by the ratio of the frequency of the time-base oscillator and the signal fed into the vertical-deflection plates.

We might take this occasion to remark that the upper limit of the horizontal-amplifier bandwidth will have a tremendous influence upon the linearity of the sawtooth time-base voltage when its source is adjusted to a relatively high frequency. This is covered in detail later in this chapter, but we cannot refrain from mentioning at this time that the useful frequency limit of the time-base oscillator is determined by the upper limit of the horizontal-amplifier bandwidth. A ratio of about 7:1 between the upper-frequency limit of the horizontal-amplifier bandwidth and the top frequency of the time-base oscillator is the smallest that can be used, still retaining some semblance of linearity in the sawtooth. This is a debatable point, we admit, because different individuals have different ideas of what departure from perfect conditions is acceptable. In this connection, we might also mention that horizontal-amplifier bandwidths invariably are quoted in terms of sine-wave amplification, and only occasionally is reference made to the fundamental frequency limits of sawtooth and square waves which may satisfactorily be passed through the amplifier. The wide-band oscilloscope specifications pay more heed to these requirements than the medium- or low-band devices. The full specifications of almost all oscilloscopes will be found in Chapter 22.

The Time-Base System

The next major system in the basic oscilloscope is the *time-base* system, also known as the *sweep* system. The latter term is commonly used to identify the origin of the horizontal-deflection voltage necessary for the production of the pattern which shows changes in amplitude of the trace with time — the waveform. In view of the detailed attention paid to the time base in the preceding chapter, it is unnecessary to dwell at any length upon its reasons for being a part of the oscilloscope.

It is, however, imperative to identify clearly the composition of the system as a whole. It is generally divisible into two parts. Referring to Fig. 10-1, one of these is the time-base *oscillator,* which is the source of the *sawtooth* voltage that causes the *linear* movement of the spot in the horizontal direction. The other part is the *sine-wave* time base which generally is used for alignment purposes rather than waveform observation. This time base was not discussed among the linear time bases in Chapter 9, but is given attention in Chapter 13. For the present we can describe it as being simply a single-frequency sine-wave voltage. The important item to bear in mind is that future references to the *time-base system* include *both* types of time-base voltages, whereas reference to the time-base or sweep *oscillator* means the sawtooth generator alone. Both types are representative of *repetitive* sweeps, in that they repeat the horizontal displacement of the electron beam until the action is deliberately stopped. The selection of either kind of time-base voltage, or rather the source of either voltage, is another function of one of the switches on the panel of the oscilloscope.

The Time-Base Oscillator (Sawtooth)

Reference to Fig. 10-1 shows that the time-base oscillator is associated with two other sections of the oscilloscope, the *synchronizing system* as well as the horizontal amplifier. But before we can explain this association between the sections, the capabilities of the time-base system deserve a few remarks.

The sawtooth-voltage generator is a variable frequency device with a sawtooth-shaped voltage output. Its general range was referred to in the preceding table, although it is true that these few examples do not give a complete cross-section of the art. But they do quote enough figures to establish that the output is variable over a range. The fact that all oscilloscopes do not employ identical sweep oscillators also is evident. Finally, the lower limit may extend to below 1.0 cycle. In this connection, it might be of interest to comment that some instruments employ sweep oscillator systems which require several seconds of elapsed time for the electron beam to complete one excursion across the screen.

There are many different ways in which the sawtooth time-base voltage may be used, but for the present, we will focus our attention on the basic function only, that is, the voltage required for the horizontal displacement of the beam to show the waveform of the quantity. The path of the sawtooth voltage for such operation is from the oscillator to the horizontal amplifier, and then to the horizontal-deflection plates of the cathode-ray tube. In some oscilloscopes of other than the basic variety, provision is made for feeding the output of the sawtooth-voltage source to an external circuit for testing or other purposes. There are two controls that are associated with the time-base or sweep oscillator on the basic device (and on most of the others as well). Both appear on the front panel. One is a multiposition switch which selects the frequency band of the oscillator. The second is continuously variable (usually a potentiometer) and is the means whereby the frequency within the range selected is accurately adjusted to the exact value required to properly display the signal fed to the vertical-deflection system.

Since the calibration of these systems is only approximate, an idea of the frequency of the signal being examined is valuable in selecting the proper setting of the range switch, although it does not require too much time to establish the proper conditions of operation even when the frequency of the vertical-amplifier signal is not known. Recognition of the required direction of change in the setting of the coarse range switch comes easily as a result of familiarity with the patterns on the screen. Details of these operations follow later in this chapter.

Sine-Wave Time-Base Voltage

The sine-wave time-base voltage, on the other hand, is a much simpler matter. As we have said before, it is a single-frequency signal and it usually is secured from the power supply as shown in Fig. 10-1. In the United States, this is 60 cps in virtually all instances. A minor departure from this value is of little consequence, since the conditions of use, as a rule, are such that the signal which is being observed also varies at the same rate as the sine-wave sweep. In some cases which do not rightfully belong in this discussion, a 120-cps signal also is made available as a sine-wave time-base voltage. This is dealt with in detail in Chapter 16.

The Synchronizing System

This part of the instrument is one of the relatively simple assemblies, although its contribution is very important. It is the means whereby the pattern on the screen is made stationary for observation. Studying a pattern is out of the question unless it is "stopped" for easy examination by the eye. Any attempt to follow the variations in amplitude of a pattern which is moving is virtually impossible, even when its motion is relatively slow.

Although represented by a simple control, usually nothing more than a potentiometer type of voltage divider (sometimes associated with special amplifiers), we nevertheless identify the entire arrangement as a system simply because it is associated with a switching arrangement which we referred to earlier and which is shown in Fig. 10-1. By means of this synchronizing arrangement, the sawtooth time-base oscillator may be "locked in step" with the frequency of the signal being fed to the vertical-deflection plates and the electron beam is made to retrace repeatedly the same path on the screen. The net result is that the pattern appears stationary to the eye.

The synchronizing process is in itself a simple one, being nothing more than the injection of a small amount of "control" voltage into the sawtooth time-base oscillator circuit, which by this means is made to oscillate at a frequency that is either the same as, or an integral submultiple of, the control-voltage frequency. Since the control voltage is either a small portion of the signal being applied to the vertical-deflection system, or is an integral multiple of that signal, it can synchronize the time-base oscillator to stay in step with the signal being examined.

Since the behavior of the sawtooth oscillator is determined by the magnitude of the control voltage injected into it, means are provided for varying the amplitude of the synchronizing signal. This is the function of the synchronizing control which is located on the panel of the oscilloscope. The remainder of the synchronizing system consists of means for securing the control voltage from sources other than the vertical amplifier, in order to satisfy different conditions of oscilloscope use. For example, if the vertical amplifier is not being used for any one of a number of reasons, the signal to be examined is fed directly into the vertical-deflection system. Synchronization of the sawtooth time base may be necessary and, in this event, the synchronizing voltage is secured from the source of the vertical signal. This source furnishes a synchronizing (control) voltage to the synchronizing amplitude control, from whence it is fed to the time-base oscillator. The purpose of the synchronizing signal amplitude control is the same as previously stated, namely, to give enough sync voltage to properly control the time-base oscillator frequency.

Still another source of the synchronizing voltage for the sawtooth time base is the 60-cps power supply inside the oscilloscope. Injecting this voltage into the time-base multivibrator will lock that system to any desired frequency which is an integral multiple of the control voltage, in this case 60 cps. In order for this to happen, the frequency setting of the sawtooth oscillator must approximate the multiple frequency. The selection of any of the three sources of the synchronizing voltage is made by means of a switch on the panel.

It is important not to confuse such 60-cps synchronization with the 60-cps time base. Synchronization is the frequency control of a sawtooth voltage; the time base remains a voltage of sawtooth shape, and may be 60 cps, 120 cps, 240 cps, or any other frequency which is an integral multiple of 60 cps. At all times the horizontal-deflection field is that developed by a sawtooth voltage. On the other hand, the 60-cps time-base voltage is fixed in frequency at 60 cps and is sinusoidal in shape. Moreover such a time-base voltage requires no synchronization, since it is definitely fixed in frequency.

The entire subject of synchronization is extremely interesting, but we will cover only the high points; a more complete discussion of the behavior of the time-base oscillator while being synchronized is given in Chapter 11.

The Intensity-Modulation System

Functionally, the intensity-modulation system, sometimes known as the *Z*-axis system, is a circuit through which a signal may be fed either to the control grid or to the cathode of the cathode-ray tube for the purpose of altering the intensity of the trace on the screen. This is done by varying the electron density of the beam; momentarily increasing the electron density increases the brilliance of the trace, while diminishing the density makes the trace momentarily disappear from the screen.

Sometimes this is accomplished by feeding the intensity-modulation signal to the cathode-ray tube through a very simple circuit, and sometimes (less frequently) through a separate amplifier which is called the Z-axis amplifier.

Intensity modulation may serve many purposes. One of these is time calibration of the trace, by which is meant a change in intensity of the trace at definite time intervals so that the separation between the altered points on the trace can be assigned time values. This makes it possible to determine the time duration of a wave, or the time lapse between points on the wave. This is a very important function in cathode-ray oscillography, although it is not a frequency application of the basic oscilloscope.

A much more commonplace function of the modulation process is the blanking of the sweep retrace. This is the pattern which develops on the screen during the time that the sweep voltage returns the electron beam to its starting point prior to making the next excursion across the screen. This will be explained and illustrated in detail later; for the present, let it suffice that the blankingf action attained through intensity modulation makes the study of waveforms much easier.

Any type of voltage, provided it has the correct polarity, can be used for intensity modulation. By this we mean that to increase the intensity of the trace a negative voltage is required for the cathode of the cathode-ray tube, and a positive voltage is required at the control grid. For beam extinction, the reverse polarity is required at the designated points. The waveshape of the modulation voltage can be almost anything from a sine wave to a sharp negative or positive pulse or a square wave. The kind of action desired on the cathode-ray-tube screen determines the shape of the voltage. The shorter the desired intervals of beam blanking or intensification, the steeper must be the voltage waveshape which affects the control grid or the cathode.

The blanking frequency is not without its limits. The R-C networks in the simple system or in the amplifier impose limits on the frequency. In fact, some Z-axis amplifiers bear bandwidth ratings which determine the range of frequencies that may be used for intensity modulation.

Another item of interest is the voltage requirements. In the simpler arrangement, amplifying means are absent in the oscilloscope. Therefore, the voltage required at the intensity-modulation input terminal is equal to that needed at the grid-cathode circuit of the cathode-ray tube. Determined by the operating conditions in the cathode-ray tube, the intensity-modulation voltage under such conditions may vary from 30 to perhaps 60 or 70 volts or more. In the absence of specific information, experimental determination is the only means of arriving at the correct level. In view of this relatively high level, and since the exact amount needed may not be conveniently available, an amplitude control at the source of the signal is advisable.

On the other hand, when an amplifier (Z-axis) is contained within the oscilloscope, as little as 1 volt at the input terminal is sufficient to intensify or blank.

The Power Supply

The function of the power supply in the cathode-ray oscilloscope is self-explanatory; it furnishes all of the operating voltages required by the tubes in the device. In the vast majority of instances, the power supply is integral to the oscilloscope, but there are exceptions to this rule. Several devices, such as the projection oscilloscope which requires high voltages, utilize a separate and individually housed power supply. Whereas the usual design of power supplies is such that their output voltages are fixed in value, the separately housed devices give variable values of post acceleration voltage, thus affording a simple means for varying the brilliance of the trace, to suit different conditions of use.

A variation in the design of these systems is found in the power transformer; the devices intended for domestic use in the United States are designed for operation at 115-125 volts on 50-60 cps power systems. Occasionally the unit is designed for use on power services with frequencies as low as 40 cps, which makes it suitable for use in foreign countries which operate power systems at this frequency. However, the suitability of the unit is still determined by the voltage rating of these services, for in many instances, the power systems in foreign nations operate at from 150 to 230 volts. The conventionally designed 115-125-volt transformer is unsuited for such application. The units intended for export have a switch located in the primary system. This device accommodates the transformer to the different voltage systems by changing the turns ratio in the transformer. So, it is easy to see that even the power supply has performance specifications.

Panel Organization of the Controls

The next phase of familiarization with the basic oscilloscope involves the various controls. Although the manufacturers of cathode-ray oscilloscopes have taken much license in the assignment of names to the various controls which appear on the panel, they have made it relatively easy for the uninitiated to use the instrument. It has become an unwritten law to place the controls for the vertical amplifier within a certain zone on

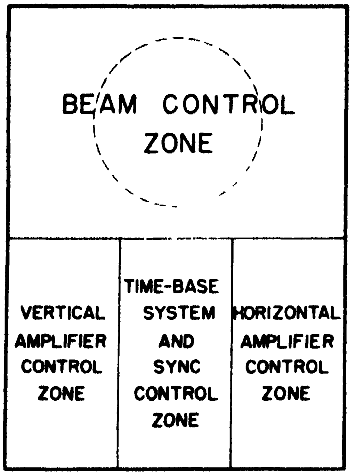

BEAM CONTROL ZONE
VERTICAL AMPLIFIER CONTROL ZONE
TIME-BASE SYSTEM AND SYNC CONTROL ZONE
HORIZONTAL AMPLIFIER CONTROL ZONE

Fig. 10-9. — Early organization of front panel controls.

the panel surface; the same goes for the horizontal amplifier, for the time-base oscillator, and even for the different controls of the electron beam.

Of course, exceptions will be found, but they are not cause for concern, because even under the most adverse circumstances, some semblance of organization based upon the old practice (which we hope will be continued) will remain. Where such deviation from standard prevails, it involves the fewer, rather than the greater, number of controls.

The very early organization of controls on the front panels of oscilloscopes resembled that shown in Fig. 10-9. The face of the panel was divided into two zones by a horizontal line. The upper one contained the face of the cathode-ray tube and the four beam controls: the intensity, focus, and vertical and horizontal centering. The lower two-thirds of the panel was divided into three zones by vertical lines. The left-hand panel contained the controls which were associated with the vertical amplifier, and which we have labeled the "vertical-amplifier control zone." Some devices had more controls than others, but those which were related to the vertical amplifier invariably were located in this area.

The right-hand panel was the zone where the horizontal-amplifier controls were located. Sometimes they were as numerous as the vertical-amplifier controls and sometimes they were not. If some other switch was located in that zone, it had some association with the horizontal amplifier. To all intents and purposes, the identity of the zone was retained.

The middle panel was the zone where, in general, the time-base system and synchronizing controls were to be found. These included the means for selecting either the sawtooth time-base voltage or the sine-wave time-base voltage and for the selection of the correct frequency of the former.

Definite lines did not indicate the boundaries of the control zones, but the impression of such zonal dis-

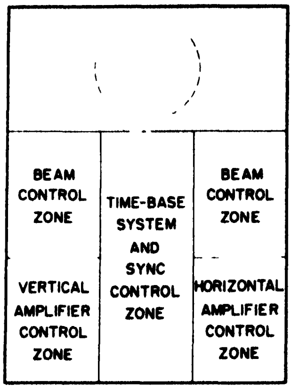

BEAM CONTROL ZONE
VERTICAL AMPLIFIER CONTROL ZONE
TIME-BASE SYSTEM AND SYNC CONTROL ZONE
BEAM CONTROL ZONE
HORIZONTAL AMPLIFIER CONTROL ZONE

FIG. 10-10

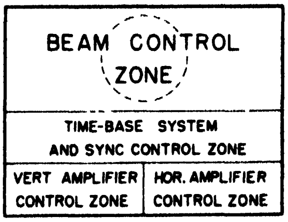

BEAM CONTROL ZONE
TIME-BASE SYSTEM AND SYNC CONTROL ZONE
VERT AMPLIFIER CONTROL ZONE
HOR. AMPLIFIER CONTROL ZONE

FIG. 10-11

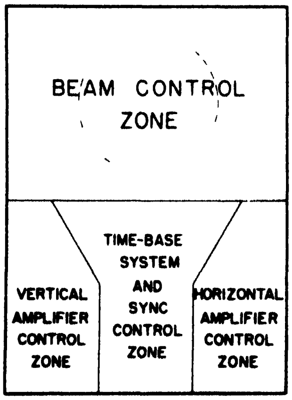

BEAM CONTROL ZONE
VERTICAL AMPLIFIER CONTROL ZONE
TIME-BASE SYSTEM AND SYNC CONTROL ZONE
HORIZONTAL AMPLIFIER CONTROL ZONE

FIG. 10-12

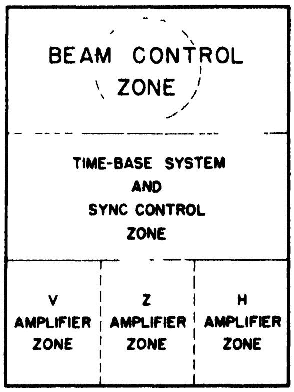


FIG. 10-13

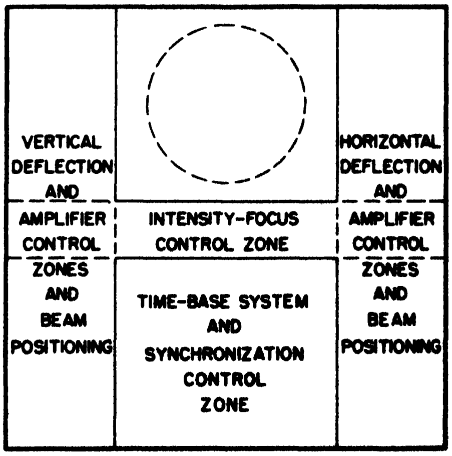

VERTICAL DEFLECTION AND AMPLIFIER CONTROL ZONES AND BEAM POSITIONING
INTENSITY-FOCUS CONTROL ZONE
TIME-BASE SYSTEM AND SYNCHRONIZATION CONTROL ZONE
HORIZONTAL DEFLECTION AND AMPLIFIER CONTROL ZONES AND BEAM POSITIONING

FIG. 10-14

Fig. 10-10 through Fig. 10-14.—Various methods of front panel control zoning used in modern oscilloscopes.

tribution of the controls was almost immediate from the first glance at the panel. That this scheme had numerous advantages was easy to see. It made familiarization with the location of the controls easier because the association with the internal circuits was definite. The association between the location of the controls and any desired change in the trace was immediate. The grouping of the controls of related character made the location of the desired control easier. It even made the labeling of the controls easier, permitting the use of abridged terms or even single letters to indicate the purpose of the control. This was especially true of the vertical- and horizontal-amplifier controls, for it allowed the use of such single words as "input" and "gain" without any ambiguity, since the location of the related control or terminal in a certain zone immediately declared its purpose in relation to the over-all system of which it was a part.

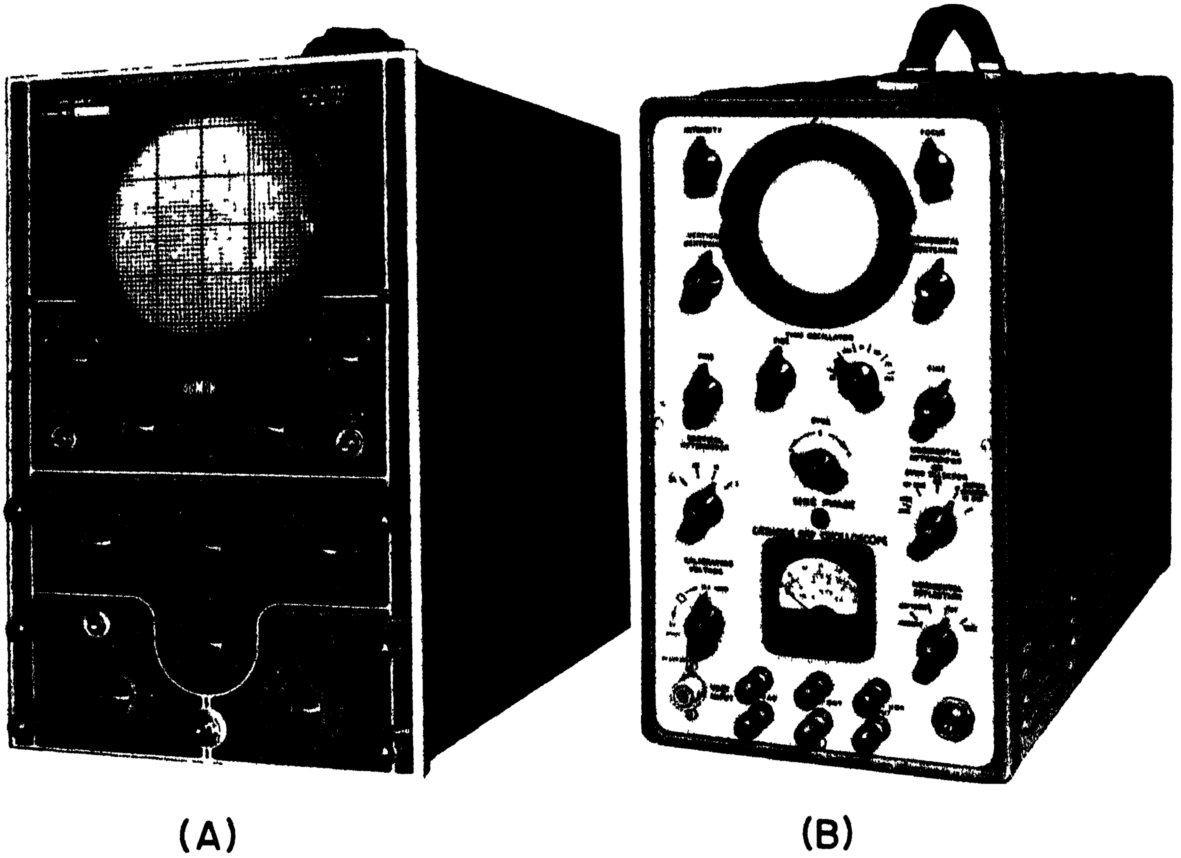
The general idea has lived on, and the modern device, especially the basic type, quite frequently follows the same plan. As would be expected, there are exceptions. Sometimes these exceptions are motivated by the desire to beautify the panel, especially when the number of controls is insufficient for uniform distribution, or when the controls are not of like physical shape and a rearrangement is necessary for esthetic purposes. Then, too, some devices are quite large, especially the more elaborate oscilloscopes, and the great number of controls required on the panel does not lend itself to such organization. In these cases, the zonal distribution of controls is maintained as best that it may be.

As we said before, the basic devices retain some form of zonal organization of the controls. As one might expect, a manufacturer will try to retain his particular form of organization on all his equipment. Some other manufacturer may have some other idea and he will put it in practice on most of the equipment which *he* makes. Fortunately, we do not have as many kinds of panel layout as we have manufacturers. All abide by the general principle, and if they desire individuality in the appearance of their devices, they try to attain it with the kind of dials, control knobs, engraving, and other auxiliary physical features which they employ. The plan shown in Fig. 10-9 is still very popular. A modification appears in Fig. 10-10. The difference is not very great, only the relocation of the beam-control elements so that they are zones with the vertical- and horizontal-amplifier controls, respectively.

Another arrangement of control zoning is shown in Fig. 10-11. This is for a panel which is wider than it is high. Still other examples of control zoning on the panel are shown in Figs. 10-12 and 10-13. Fig. 10-13 is a large unit and the panel carries quite a number of controls. The panel is divided horizontally into three areas and the bottom zone is divided vertically into three areas, each area being devoted to the systems identified in the illustration. Finally, another method of control zoning is shown in Fig. 10-14. This, too, is one of the very elaborate oscilloscopes. There are three vertical panels. The left panel contains all of the controls which have any bearing on the vertical motion, including the positioning of the beam, the vertical amplifier, terminals for direct connection to the vertical-deflection plates and also the intensity-modulation system controls. The right-hand panel contains all the corresponding controls for the horizontal motion, and the center zone contains all the controls relating to the time-base voltage and synchronization. The narrow central horizontal panel might be considered the location of the beam controls, with the vertical- and horizontal-positioning controls being appropriately located within the vertical- and horizontal-deflection control zones.

These few examples of control layout on the panels of oscilloscopes are quite representative of those which will be found today. Sometimes the organization of the controls is indicated by grouping of the controls with functional labels which cover a group of individual, but related, controls. An example of this is shown in Fig. 10-15A. Sometimes a title indicating the functional relationship of a group of controls appears on the panel, and the identification of the controls under that heading or above it is done in the simplest manner, because it is felt that the association of ideas will convey the necessary information. This appears in Fig. 10-15A, for example under the headings of X-AXIS and Y-AXIS, and also in Fig. 10-15B, where the term VERTICAL ATTENUATOR above the step attenuator also indicates the purpose of the continuously variable potentiometer located directly above it and labeled by the single word FINE. The same is true of the group of controls related to horizontal deflection, where the words HORIZONTAL ATTENUATOR above the switch help define the function of the control directly above it labeled FINE. In this connection, we might also mention the label TIME OSCILLATOR above a variable control and a switch which apparently selects frequency ranges. The meaning of the word FINE above the uncalibrated control knob is obvious, but, again, only because of the association of ideas.

All of this leads us to a very important part of the study of the controls and their functions. It is the nomenclature used with the controls and the labels on the face of the oscilloscope.



(A) Courtesy Du Mont Labs (B) Courtesy RCA

Fig. 10-15—Two typical oscilloscopes illustrating the functional grouping of controls.

Nomenclature of Controls

The absence of standardization of nomenclature among cathode-ray-oscilloscope manufacturers has resulted in the use of a wide variety of names for identification of the controls, switches, and other components. Quite frequently, identical elements are specified by different names and dissimilar controls by like names, on different instruments. At first glance, this situation may seem to be quite a problem for the individual who is first acquiring a familiarity with cathode-ray devices. It is not as bad as it seems. The amount of confusion is kept to a minimum provided that certain fundamental facts are recognized. The first of these is the general plan of panel organization; related controls are near each other, or are at least located in a group. Second is the use of similar terminology for the basic sections of the oscilloscope. When these two features are correlated, the general meaning of the labels becomes clear. Even the use of two entirely different labels for the same control will not prevent comprehension of the function of the control.

For example, such words as VERTICAL or VERT, or single letters such as V, or the mathematical expression Y, which denotes the vertical axis when applied to oscilloscopes, all have the same meaning. The control, switch, or terminals that bear one of these labels are immediately associated with the motion of the electron beam in the vertical direction. Such motion may be due to the signal, or it may be due to an operating voltage.

As examples let us consider the following: the words VERTICAL or VERT, or V or Y ahead of the word POSITIONING or GAIN or INPUT, immediately denote the system in which the labeled control or switch or terminal plays its part, and also the function of the component. Words such as HORIZONTAL or HOR, or letters such as H or X, all indicate the same thing, the horizontal-deflection channel, and the word which follows these prefixes serves as the statement of the function of the component.

The similarity between the vertical channel and the horizontal channel is very great, although the exact performance specifications of these channels may not be the same. Many components in each perform like functions, and consequently bear analogous labels. Y GAIN and H GAIN, or VERT POSITION and HOR POSITION, or V INPUT and H INPUT are examples.

Operating Control Nomenclature

An understanding of the distinction between the operating voltage controls and signal controls will be helpful in making the panel of the basic oscilloscope more readable. This is important in the light of the profusion of labels which one may find on the faces of these devices.

Several controls are related to the no-signal condition of the cathode-ray tube. These controls establish the electron beam required for proper *viewing* or proper *display*. We call these the *operating* controls.

The operating controls are the intensity, the focus, and the two beam-position controls. *All the rest* on the face of the panel are *signal controls,* or components which are related to the signal in one way or another. Having set the operating controls apart from the signal controls, we are ready to consider the different names that are assigned to them.

Intensity, Focus, and Positioning Control Nomenclature

It would hardly seem possible that names such as these, which indicate functions so directly, can be subject to such variation. Yet this is the case, with one exception, the focus control. The remainder often appear with different names. Each refers to some form of control of the beam by means of *operating voltages only*. We emphasize this point in order to prevent the confusion of the intensity control, which determines the brilliance of the trace, with intensity modulation, a signal control.

The names associated with the operating controls do not differ greatly in general. With the exception of the intensity control, there is little chance of misunderstanding the meaning of the labels. Although a variety of words are used to express a function, all are synonymous, although the very exacting individual might contend that one term more completely describes the action than another. The following is a list of the names which may be found on oscilloscope panels as identifications of the four operating controls.

Control Function	Names Assigned
Beam (Trace) Intensity	Intensity Brilliance
Beam Focus	Focus
Position of Beam (Trace) along Vertical Axis	Vertical Position Vert Position Vertical Centering Y Position Y Centering V Centering V Position Positioning
Position of Beam (Trace) along Horizontal Axis	Horizontal Position Hor Position Hor Centering H Position H Centering X Position X Centering Positioning

These names need little discussion. If any comment is justified, it is relative to the use of the word POSITIONING without any reference to direction. This is made possible by the placement of the control within that zone on the panel where some other reference to the axis is made, thus immediately correlating the positioning control with its axis.

Two methods are used in basic oscilloscopes for beam positioning. One of these employs a single potentiometer in each of the positioning circuits, and the other utilizes a dual potentiometer. These systems are shown as (A) and (B) in Fig. 10-16. The simpler of the two is shown as circuit (A). By suitable connection to the power-supply system, a fixed d-c voltage is arranged across two potentiometers, *R1* and *R2*. The moving arm of potentiometer *R1* is connected to one of the horizontal plates, and the moving arm of potentiometer *R2* is connected to one of the vertical plates. The remaining horizontal and vertical plates are grounded. Thus the moving arm of each potentiometer can apply a variable d-c voltage to each of the deflection systems, making one plate of each pair positive or negative with respect to its associated plate. This moves the electron beam in the appropriate direction. By varying the settings of the potentiometer arms, the static position of the beam is controllable over a substantial range. The signal, on the other hand, is applied to each deflection system via the two capacitors *C1* and *C2*. Since these are a-c paths, they do not affect the static settings of the positioning fields. When a signal is applied, the signal voltages are superimposed on the static positioning voltages and produce the resultant fields that develop the trace on the screen.

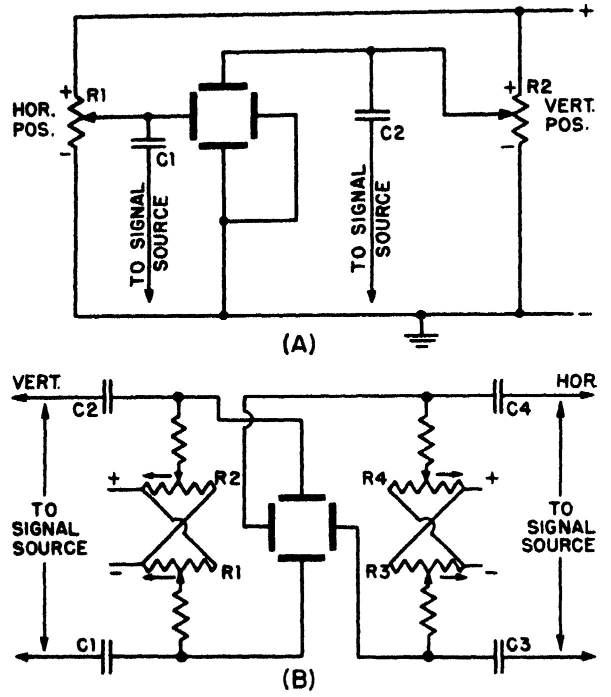


Fig. 10-16. — Two beam-positioning systems, (A) unbalanced, (B) balanced.

Such a positioning system varies the potential of one plate relative to the other, and will be found used in those oscilloscopes which do not utilize push-pull amplifiers.

The second system used in many basic oscilloscopes is shown in Fig. 10-16B. In (B) the potentials of both of the deflection plates in each pair are varied. Each positioning system consists of two potentiometers, *R1-R2* and *R3-R4*, respectively. Let us analyze one of these. Suppose we select the vertical-deflection system. The two potentiometers *R1* and *R2* are cross-connected electrically and mounted on a single shaft so that the movable arms are manipulated simultaneously. It is important to remember that both arms always move in the same direction.

By virtue of the cross connections of the two resistance elements and the manner in which the d-c voltage secured from the power supply is supplied to the two potentiometers, the opposite ends of each resistance element are positive and negative, respectively. Zero voltage prevails at the midpoint of each arm. Therefore, as the arms are moved in either direction from the midpoint, the voltage applied to one of the deflection plates will be of some value and one polarity and the voltage applied to the other deflection plate will be of equal value but of opposite polarity. Thus, when one deflection plate is made 20 volts positive, the other is made 20 volts negative, and both act on the beam. This exemplifies balanced positioning. When the two arms are at the midpoint of the resistance elements, zero-positioning voltage is applied to the two plates and the beam occupies the midpoint between the two plates. By moving the arm toward the right or toward the left, the beam is moved upward or downward.

The same reasoning applies to the horizontal-positioning system except that, in this case, the axis of beam motion is horizontal, whereas, in the previous case, it was vertical, relative to the screen surface.

The capacitors *C1-C2* and *C3-C4* are the isolating capacitors in the signal input circuits. They prevent the d-c voltage on the plates of the output tubes from affecting the static potentials of the deflection plates. The signal voltage from these sources is superimposed on the d-c potentials from the positioning systems.

Still another positioning system is used in some instruments, but since this is found mostly in oscilloscopes which contain direct-coupled amplifiers, and since these are not representative of the basic variety, we shall hold their descriptions in abeyance until later in this chapter.

Vertical-Amplifier Control Nomenclature

Relative to nomenclature associated with the vertical-deflection amplifier section, we must introduce several circuit points which we did not discuss during the general discussion of the section functions. In Fig. 10-17A and B (which is a segment of Fig. 10-1) is shown an elaboration of the vertical-deflection amplifier section. The controls usually associated with this part of the basic oscilloscope have been added.

The apparent simplicity of the system shown in (A) in comparison with (B) stems from the difference in the controls which may be found in two classifications of basic devices. We can speak about these now since we are delving deeper into the construction of these units. In other words, a variety of instruments may be representative of the basic variety, yet they need not be identical in every respect. Fortunately, certain dividing lines exist between these basic units.

This is evident from (A) and (B) of Fig. 10-17. In the former, the gain control is a simple potentiometer type of attenuator R_V and the input to the amplifier is via two binding posts, which, in some cases, may be replaced by two pin-jacks. In (B) of the same figure, the amplifier signal attenuator consists of two controls, a step attenuator R_S and the continuously variable potentiometer R_V. The means of getting the signal into the amplifier is via either of two systems, the two binding posts or a probe connector, the selection of either of these two means being made by a two-position switch.

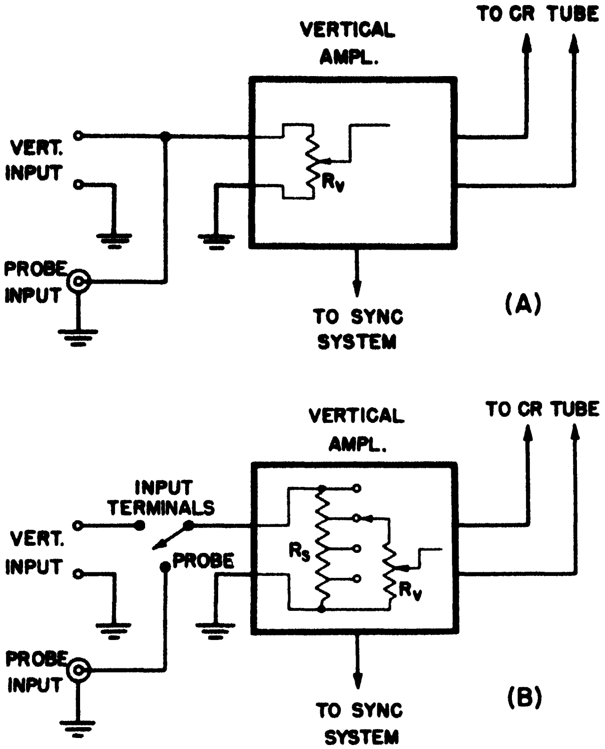


Fig. 10-17. — Two methods of vertical gain control; (A) using a simple potentiometer and (B) using a step voltage divider in combination with a potentiometer.

We hasten to mention that the use of a dual-section input attenuator does not automatically mean that probe input as well as binding-post input will be found in the instrument.

As is evident from an examination of (A) and (B) of Fig. 10-17, this difference in design of the input system of the vertical amplifier has a major effect on the number of controls which appear on the panel of the instrument, which are related to the vertical amplifier. In addition, as will be seen, this difference in design has a major effect on the labels selected for the identification of the different controls.

In order to avoid duplication of effort in the listing of the control names, we have elected to use the type of signal-attenuation system present in the input to the vertical amplifier as a dividing line between the two kinds of basic devices and the names assigned to the controls. An examination of the complete schematic wiring diagram of an instrument (such as may be found in Chapter 22) will disclose the exact circuits of the amplifier input system and indicate whether the device is in category (A) or (B).

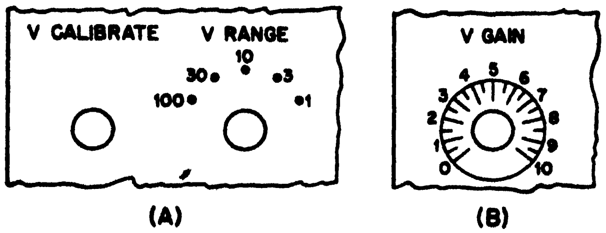
It is possible that the general pattern of circuits can be determined from an examination of the panel, since there is a distinct difference in the number of controls associated with the vertical amplifier when the input systems to the vertical amplifiers differ. As a matter of fact, the presence of calibration on one of the controls is a clue to the organization of the gain-control system, whether it conforms with (A) or (B) of Fig. 10-17.

In (A) of Fig. 10-17, three elements are related to the vertical amplifier, these being the input binding posts, the gain control, and, infrequently, the probe connector. The organization of (B) (Fig. 10-17), on the other hand, involves five items. These are the input binding posts, the probe connector input, the terminals-probe selector switch, the step attenuator R_s, and the continuously variable control R_v. Comparing these two, there is very little distinction between the input binding posts on each, or in the probe connector input. In other words, these might not afford definite identification of circuit makeup. The terminals-probe selector switch constitutes a clue, however, since it is a practice to provide this facility only in units which make use of the dual-section gain controls. This is true, not because the design of the gain control determines what shall be the design of the signal input system, but rather because the class of instrument which employs a combination step and continuously variable attenuator usually dictates that the probe input be apart from the binding-post or pin-jack input, thereby permitting each to be used independently of the other.

More positive identification of a system which conforms with (B) rather than (A) of Fig. 10-17 is the markings on the dial or the panel for the step attenuator. Although the labeling of the control is not a sure sign that it is a multiposition switch, momentary manipulation of the knob will confirm the implication of the markings. What we mean is that, physically speaking, this is a multiposition switch; such construction is instantly detectable upon manipulation, but even before being handled, such calibration as 1, 10, 100, and 1,000, or decimal values such as 0.1, 0.01, and 0.001 in discrete steps is a sign of its purpose as a decade attenuator. The two scales mean the same thing in the final analysis. The former type of numbering expresses the ratio between the input and output voltages as, for example, 1:1, 10:1, or 100:1, etc., whereas the other kind of scale expresses the output voltage in terms of fraction of the input voltage. Examples of these are unity transfer or 1:1, and one-one hundredths of the input voltage at position 0.01, etc.

Before concluding this discussion concerning the step attenuator, it would not be out of order to refer to another physical variety, although it is not to be found among the examples of basic oscilloscopes. We make these statements here because they fit the organization of the text well. We are referring to the step attenuator which has push-button construction, such as the push buttons which are used to select stations on the conventional radio receivers. Such construction is used occasionally to replace the rotary switch. Each of the buttons bears the markings which would normally be found at the various positions of the rotary switch. The fact that physically the switch is of different construction does not alter its utility, nor does it imply any other meanings for the attenuation factors which are the labels on the buttons.

A step attenuator of the dual type is never used without the continuously variable potentiometer; the latter is necessary if continuous variation of the signal is to be attained. As a rule, the range of signal-level control available with the continuously variable potentiometer is 10:1, which fills in the gaps between steps of the step control. The two units acting in concert, therefore, afford continuous control of signal level from unity to a maximum of 1,000 or 10,000, depending on the individual design. An example of a panel bearing a step and a continuously variable attenuator control is shown in Fig. 10-18A. In this particular instance, calibration of the continuously variable control is absent; usually some sort of scale markings are available for reference. In either case, the usual range of attenuation attained with the variable control is 10:1 and this is also the case when the graduations on the scale are from 1 to 100 instead of 1 to 10. Concerning the simple arrangements in (A) of Fig. 10-17 very little need be said. It is simply a potentiometer-type gain control, and while the scale graduation (if one exists on the panel) may be like the calibrated device, it remains simple linear attenuation and is not rated at a fixed amount of attenuation. A typical control and its scale is shown in Fig. 10-18B.



After RCA

Fig. 10-18.—Front panel appearances of vertical gain controls.

Relative to nomenclature, the organization of Fig. 10-17A and B introduces possible confusion only in connection with the gain controls. There can be no mistaking the signal input points, be they binding posts, pin-jacks, or the probe-connector. These are positive and unmistakable, yet it is possible for the completely inexperienced to err simply because of the multiplicity of names which may be found assigned to the terminals. We shall, therefore, list these names.

INPUT TERMINAL IDENTIFICATION

Function	Names Assigned
Feed Signal to V Amplifier (Terminals or Pin-jacks)	Vertical Input
	V Input
	Y Signal Input
	Input
	Vertical
Probe	Probe
Select Probe or Terminal Input	Terminals — Probe

The probe input label needs no comment even if it is just a single word. The lack of correlation with the vertical amplifier is attributable to first, its usual location in the zone where the vertical-amplifier controls are located, and second, the fact that *only the vertical amplifier is equipped with probe input.* The latter status allows it to be placed in any location on the panel, whatever position is most advantageous to use.

As to the input terminals, a slight amount of elaboration is required. Sometimes, although not too often, one of these terminals may be marked "High" and the other may be marked G or GND. These last two markings are quite commonplace. The word "High" indicates the high side of the circuit, electrically speaking, and G or GND indicates that the terminal is at ground potential in the circuit fed through the binding posts. We might mention that when such markings are missing from input terminals which are arranged one above the other, the upper terminal usually is "High" and the lower is ground. When the terminals are side by side horizontally and unmarked, the left terminal is "High" and the right terminal is ground.

Concerning the use of the general term INPUT, again it is a matter of association of ideas. Being located in the zone of vertical-amplifier controls, license is taken to omit the direction of deflection. In the case of the word VERT adjacent to the input binding position without any functional reference, appreciation of the purpose of binding posts is necessary, if the instructions contained in the manual which accompanies the equipment are unavailable.

The identification of the terminals-probe switch is so obvious that further comment is unnecessary.

The names employed to identify the input signal controls are quite varied. Some are more appropriate than others and it is hoped that some day standardization will prevail. Assorting these names in accordance with the two input control systems we discussed previously, they are as follows:

SINGLE-ELEMENT CONTROL OF SIGNAL INPUT LEVEL

CONTROL	NAMES ASSIGNED
Potentiometer R_v (Fig. 10-17A)	Attenuator
	Vertical Gain
	Vertical Amplifier
	V Amplifier
	Y Amplitude
	Vert Amplitude
	Gain

Once more the ability to use a single word such as GAIN without any additional reference to the system which it serves stems from its location in the vertical-amplifier zone on the panel.

DUAL-ELEMENT CONTROL OF SIGNAL INPUT LEVEL

CONTROL	NAMES ASSIGNED
Step Voltage Divider R_s (Fig. 1017B)	Attenuator
	Vertical Gain
	Vertical Range
	Amp Ratio
	Attenuation
	Signal Atten
	V Sensitivity
Potentiometer R_v (Fig. 10-17B)	V Vernier
	Y Amplitude
	Vert Gain Vernier
	Y Gain
	V Gain
	V Calibration
	Gain

In the light of the varied names which appear in the above listing, it would be advantageous if the manufacturers of cathode-ray oscilloscopes chose two names from this listing and made them standard for these controls. Reference to gain as a name for these controls is not strictly correct, whereas reference to amplitude is, so why not assign a name such as "Amplitude Ratio" to the step control and simply "Amplitude" or "Amplitude Vernier" to the continuously variable element. Each of these names could be preceded by the letter V or Y to show its channel of operation.

The zonal distribution of controls has also resulted in the occasional omission of any form of identification for one or the other of these elements. As obvious as it may be to the designer and as familiar as each control may eventually become to the constant user, the absence of identifying labels imposes an unnecessary hardship, especially during the early days of familiarization.

Horizontal-Amplifier Control Nomenclature

When examining panel elements representative of the horizontal-deflection amplifier, we note a distinct parallel to the vertical-deflection amplifier. Since both are amplifiers with fairly similar duties, such a state is not surprising. By and large, the horizontal-amplifier controls found on the front panel of any oscilloscope consist of the input terminals, one or two controls used for setting the level of the input signal, and a multiposition switch which determines the availability of the horizontal amplifier for signals which issue from devices external to the oscilloscope and for the selection of different signals from within the instrument. This switch is common to all oscilloscopes, regardless of the signal-level control arrangements, indicated in Fig. 10-19A and B.

The names assigned to these controls in our symbolic illustrations are arbitrary, especially the switch we call the "Signal Input Selector." It so happens that the input terminal designations conform with some practices, only because such a label is obvious. Relative to the signal-level controls R_s and R_v, their identifications bear names which are similar to those given their vertical-amplifier counterparts, differing only in the first word which associates the elements with the horizontal rather than the vertical amplifier.

The control which deserves the most attention is the signal input selector. As indicated in Fig. 10-19, it affords a means for connecting the horizontal amplifier to the sawtooth time-base oscillator, the 60-cps sinewave time-base voltage source, or some external signal source. These respective switch portions are usually labeled as shown, namely INT, 60~, and EXT. Considering its duties, it is not unreasonable that it bear some label which involves the horizontal amplifier and also the sweep systems, as for example HOR/SWEEP SELECTOR, which happens to be a name that is employed by several manufacturers.

Sad to relate, however, the situation is not as simple as the above might indicate. The number of contacts on this multiportion selector switch is not always limited to the three we have illustrated. Quite frequently other functions are made part of the duties of the switch. For example, the step attenuator R_s may be a part of the

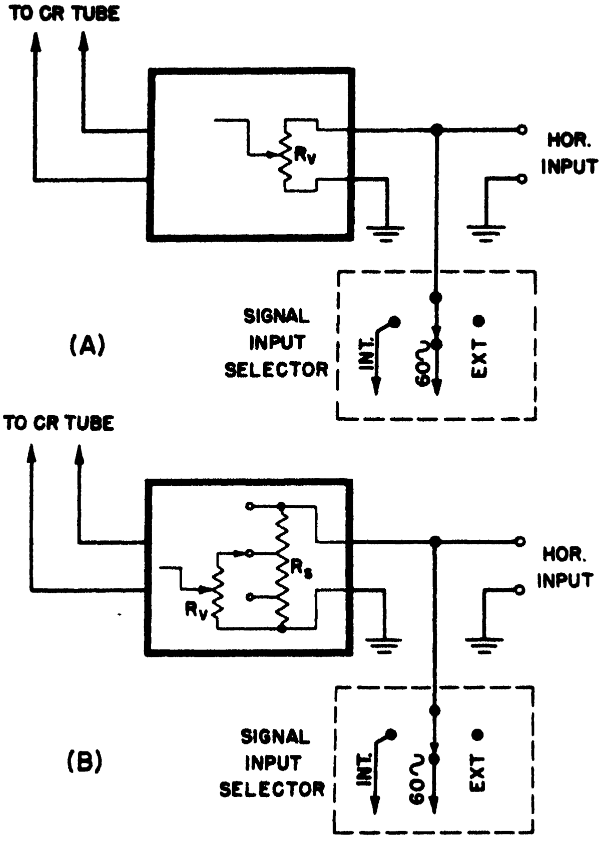


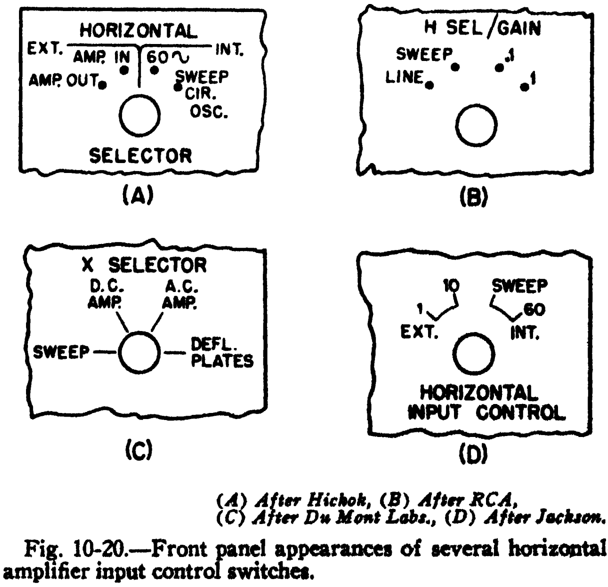
Fig. 10-19.—A typical signal input selector switch with different attenuation systems in (A) and (B).

switch. By restricting the attenuation factor to a maximum of 10, and by designing the time-base systems in such a fashion that the continuously variable potentiometer R_v alone can adequately control the signal levels from the two time-base voltage sources, the step attenuator occupies two portions of the switch and it is in the circuit only when the horizontal amplifier receives its signals from some source external to the oscilloscope. Arrangements of this type can well demand a label which differs from the first sample we stated, since the functions of the switch have been expanded. It might well bear a name which associates it with attenuator action, as for example, H SEL-GAIN or even such an adequately simple title as HORIZONTAL INPUT CONTROL.

While on the subject of attenuators, we might mention that step attenuators used in the horizontal-amplifier channel do not as a rule duplicate the vertical channel, that is, in the number of steps, or the range of signal attenuation, but sometimes they are identical, including the push-button variety. In most instances, the degree of attenuation in the horizontal channel is limited to two steps, unity transfer, or 1 :1, and 10-time reduction. This gives rise to the use of a toggle switch as the step attenuator in place of a rotary switch. The appearance of several horizontal-amplifier input control switches with and without attenuator positions are shown in Fig. 10-20. Relative to these illustrations, it might be well to comment that the word external (EXT) sometimes is labeled AMP OUT and that the word (INT) is sometimes shown as SWEEP.

It is interesting to note that not only does the nomenclature assigned to this horizontal-amplifier input signal switch differ, but the specific design relative to the circuits under control also differs, so much so that it is virtually impossible to make broad statements concerning its general appearance on equipment. There are almost as many examples of its absence as there are examples of its use. Relative to the former arrangement, it does not mean that control of these different circuits is lacking (they are available because they are necessary), but it does mean that, in some instances, the selection of the circuits is divided among a number of other switches which are part of other sections. These will appear when we deal with the time-base oscillator and with the synchronizing sections of the basic oscilloscope.

Any attempt to label these switches is beset by these limitations. For this reason, the names which are contained in the list of control identifications are those most commonly found on oscilloscope panels. Those which would demand a listing in terms of other complete sections have been omitted, because they are referred to later when the remaining controls on the panel are analyzed. Concerning the labeling of the panel controls, the order of presentation will be predicated on the fact that when step attenuators are used,



(A) After Hickok, (B) After RCA, (C) After Du Mont Labs., (D) After Jackson.

Fig. 10-20.—Front panel appearances of several horizontal-amplifier input control switches.

they invariably are part of the horizontal-amplifier input selector switch. Therefore, the continuously variable (potentiometer) gain-control list will encompass the names used for this control for both single- and dual-section arrangements.

The input terminals, too, are treated collectively, since their names do not change much with equipment design.

H INPUT TERMINAL IDENTIFICATION

FUNCTION	NAME ASSIGNED
Feed Signal to H Amplifier (Terminals or pin-jacks)	Horizontal Input H Input X Signal Input X Input I Input Horizontal

Relative to additional markings such as HIGH and GND which may appear adjacent to the individual terminals, the comment concerning the vertical amplifier applies. As to the functional labels without reference to the axis, such as INPUT, the zonal organizations of the panel develop the necessary associations with the related axis.

H AMPLIFIER INPUT SWITCH IDENTIFICATIONS

FUNCTION	NAME ASSIGNED
Control Signal Input to H Amplifier	Hor Gain/Sel Hor Sel/Gain Horizontal Sync H Sync/Sel H Sync/Sweep Sel

Several names apply to almost identical controls, whereas some of the others are appreciably different. Now there are the oscilloscopes which do not use any labels for this switch. These simply locate it among the horizontal-amplifier controls in a certain area on the panel and let it go at that. This is not as bad as it seems because, in the final analysis, the manufacturer's instruction manual for a specific instrument explains the functions of each control.

Horizontal-Amplifier Selector Switch Position Nomenclature

A few words about the nomenclature assigned to the position on this switch may be of interest, especially when different labels are used for the same thing. We have commented on the interchange of whole numbers and decimal numbers as indications of attenuation levels. The use of either is equally effective.

At least five different labels are used to indicate the position which selects the sawtooth oscillator contained within the oscilloscope. These are:

1. INT
2. Sweep
3. Sawtooth
4. S. S. Oscillator
5. Amp In.

The sine-wave time-base voltage position usually is labeled in one of three ways:

1. Line
2. Line Frequency
3. 60~

and finally, the position which makes the horizontal amplifier available for signals originating external to the oscilloscope may be found labeled as either

1. EXT
2. Amp Out.

Examples of almost all of these markings will be found in Fig. 10-20.

Time-Base Control Nomenclature

The description of the time-base controls used in the basic oscilloscope is relatively simple. This does not mean that standardization has crept in unnoticed, but rather that the use of two controls for the repetitive sawtooth sweep has become standard practice because of the circuit design. Under the circumstances, there is little likelihood of mistaken identity, although, as you will soon see, the number of different names which are assigned to these controls are not too few.

While the time-base system in the basic oscilloscope may be tied into a number of sections so that other section controls also determine the active presence of either the sawtooth oscillator or the sine-wave time-base voltage source, only the former has its own variable controls. One of these is a multiposition rotary switch and the other is a continuously variable element of the variable resistance kind. These two controls enable the selection of the exact sawtooth voltage frequency necessary for the presentation of the desired number of cycles of the signal under investigation on the screen of the cathode-ray tube.

When symbolized schematically as in Fig. 10-21, which is an abridged version of the conventional sawtooth oscillator, the switch selects the value of capaci-

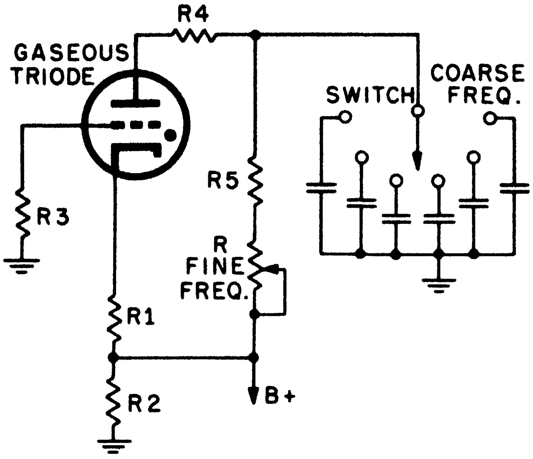
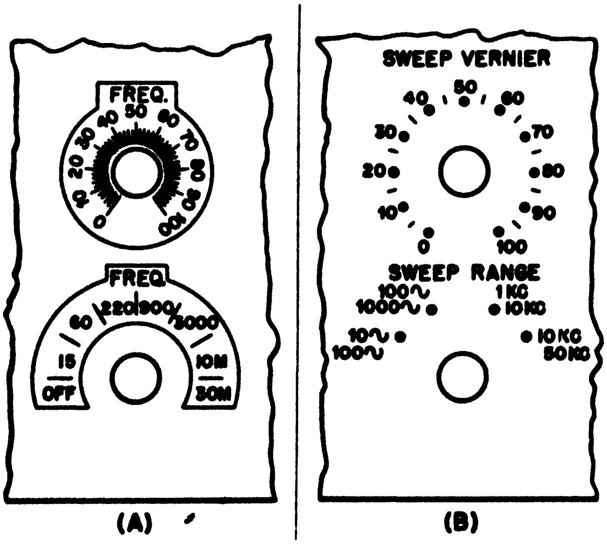


Fig. 10-21.—Schematic of a typical sawtooth oscillator showing coarse and fine frequency controls.

tance which will be charged and discharged during the operational cycle. This is the "coarse" control of frequency. The variable resistor *R* in the same illustration determines the rate at which the capacitor selected by the switch will be charged. The faster the rate of charge (assuming a fixed voltage level at which the capacitor will discharge) the higher is the frequency generated by the oscillator. This variable resistor is the "fine" frequency control.



(A) After Du Mont Labs., (B) After Television Equipment Corp.

Fig. 10-22.—Front panel appearances of sawtooth oscillator frequency controls.

The appearance of these two controls on the panel of two typical oscilloscopes is shown in Fig. 10-22A and B. The difference between the two is found in the frequency calibrations of the switch which determines the approximate settings. In (A), each position of the switch bears a frequency label and the range of frequencies which may be secured by rotation of the continuously variable frequency control is that between the value indicated for the position at which the switch is set and the adjacent one on the lower-frequency side. Thus if the switch is set at 900 cps and the next lower position is marked 220 cps, the frequencies available by manipulation of the continuously variable frequency control will be approximately 220 to 900 cps.

This is shown more clearly in Fig. 10-22B, where the calibrations of the frequency-range switch are in terms of the frequency range available at the individual switch positions. This time, too, the variation in frequency is accomplished by manipulation of the continuously variable control. Thus if the switch is set at the position marked 100-1,000, it means that manipulation of the continuously variable control will adjust the time-base oscillator so as to produce any desired frequency between 100 and 1,000 cps. The same reasoning applies to all of the other switch positions and the frequency-range markings which they bear.

In view of the presence of other kinds of time-base voltage sources in oscilloscopes, even though we are concerned only with the repetitive sawtooth variety at the moment, it would do no harm to comment that the controls we have described will be found to apply equally well to those oscilloscopes which contain the means for triggering the time-base oscillator system into operation. This is not commonplace in the devices which we consider to be basic in character, but it is an appropriate place in the text to mention it, although the subject receives separate treatment later in this chapter.

Concerning names assigned to these systems, it is interesting to note that the main heading of the system as a whole is subject to variations. We have used two names in referring to the source of the time displacement signal: the time-base oscillator and the sweep oscillator. We might add two other names which have enjoyed use, namely "time oscillator" and the single word "timing." The meanings are the same as of the other two titles. Actually the first three of these four names are in use today. Reference to the system as "timing" was employed on oscilloscopes made prior to 1940 and we recognize it today only because those prewar devices may still be in use.

As to the controls, the following is the list of names which will be encountered on oscilloscope panels. Only those which bear some relation to the repetitive sawtooth voltage sources are given.

TIME-BASE OSCILLATOR CONTROL IDENTIFICATIONS

FUNCTION	NAMES ASSIGNED
Selects Frequency Range (Switch in Fig. 10-21)	Steps Coarse Frequency Frequency Range Sweep Range Range Switch Sweep Frequency Range
Accurate Adjustment of Frequency (*R* in Fig. 10-21)	Sweep Vernier Fine Frequency Vernier Frequency Vernier Frequency Fine

Although a variety of names are used, it is clear that with a few exceptions, some of which are found in the older units only, the choice of titles has been such as to pin-point the purpose of the control. The examples which are the exceptions are such names as "steps" for the rough approximation of the frequency and "frequency" for the accurate adjustment control. These could have been selected with greater care.

Synchronizing System Control Nomenclature

As we stated earlier, the synchronizing system is very closely related to the time-base oscillator, for it is the means of keeping the time-base voltage variations in step with the amplitude variations of the signal being examined, in that way keeping the signal pattern stationary on the screen. In fact, we can go so far as to say that the synchronizing system is an adjunct of the time-base oscillator circuit, even though the two systems may be physically located at different points in the device.

When symbolized schematically, the simple synchronizing system appears as in Fig. 10-23. This circuit is essentially that of the elementary time-base oscillator to which has been added the capacitor *C* and the potentiometer *R*. This series *R*-*C* combination is the path of the synchronizing voltage (whatever may be its origin) into the sawtooth sweep oscillator. The magnitude of the "locking" voltage is determined by the position of the sliding arm of the potentiometer, so that the control is of the continuously variable type. The calibrations on such a device are arbitrary and the scale usually is graduated in divisions which are numbered from 1 to 100. The appearance of such a scale is shown in Fig. 10-24A.

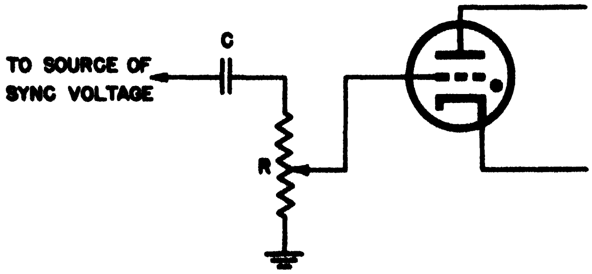
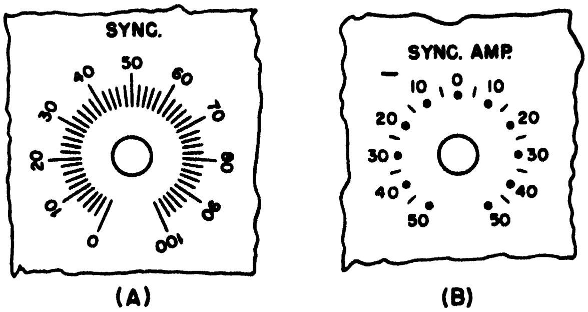


Fig. 10-23.—Schematic of a simple sync control system.



(A) After Du Mont Labs., (B) After Television Equipment Corp.

Fig. 10-24.—Front panel appearances of synchronizing controls.

Another variety of synchronizing system is somewhat more elaborate, although it will be found in basic oscilloscopes. It is of interest to us because of the difference in the calibrations used by the control, which we shall show in a moment. When symbolized schematically, the circuit appears like that shown in Fig. 10-25, where an amplifier stage, or rather a phase-splitter stage, exists between the source of the synchronizing voltage and the time-base oscillator. It is possible that additional stages may be found between the phase-splitter input and the time-base oscillator, but this is unimportant at the moment. The purpose of the phase-splitter stage is to make available at its output a signal voltage which is either positive or negative relative to ground, thus determining the instant (relative

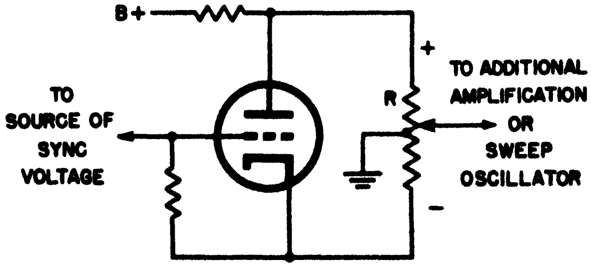

B+
TO SOURCE OF SYNC VOLTAGE
R
+
–
TO ADDITIONAL AMPLIFICATION OR SWEEP OSCILLATOR

Fig. 10-25.—Schematic of a synchronizing system using a phase splitter.

to phase) when the sweep voltage will be locked with the signal voltage. This is fully explained in Chapter 11, hence we need discuss only with those points which involve our subject, the control.

This voltage appears across the center-tap-grounded potentiometer *R* and gives rise to the plus-zero-minus form of calibration of the control. The numbers on the scale of the control advance upward from each side of the 0 at the center, each side representing a polarity zone. As the control is advanced, the signal voltage output advances substantially linearly. The appearance of such a scale is shown in (B) of Fig. 10-24. The split potentiometer also is a continuously variable control, but as far as the synchronizing signal is concerned, the range of voltage variation for any one polarity of signal exists along only one-half of the full arc of rotation of the control knob.

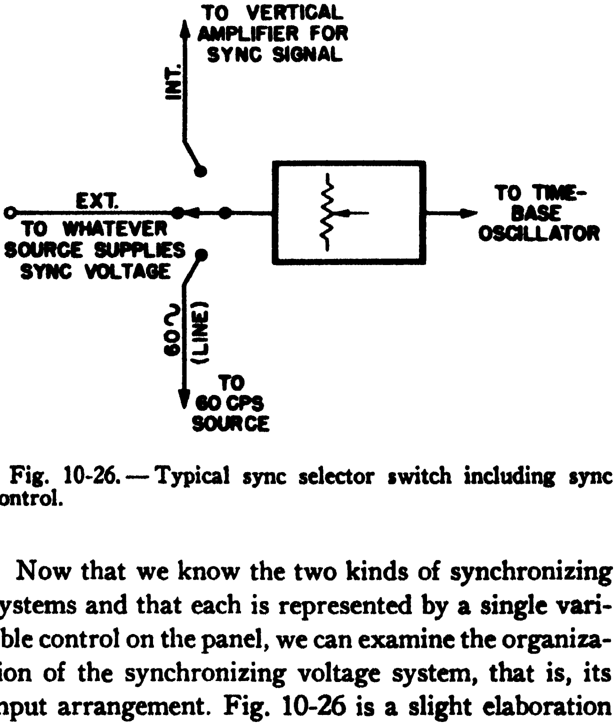

TO VERTICAL AMPLIFIER FOR SYNC SIGNAL
INT.
EXT.
TO WHATEVER SOURCE SUPPLIES SYNC VOLTAGE
TO TIME-BASE OSCILLATOR
60 ~
(LINE)
TO 60 CPS SOURCE

Fig. 10-26. — Typical sync selector switch including sync control.

Now that we know the two kinds of synchronizing systems and that each is represented by a single variable control on the panel, we can examine the organization of the synchronizing voltage system, that is, its input arrangement. Fig. 10-26 is a slight elaboration of Fig. 10-1 in that the origin of the various synchronizing signals is shown, as well as the presence of the potentiometer type of control. Concerning the controls usually found on the panel and labeled relative to the synchronizing voltage, there are usually two, the multiposition switch which selects the origin of the synchronizing voltage and the potentiometer which we have described.

Relative to the selector switch, its action is fairly obvious, as are the sources of the synchronizing voltage. The latter may issue from the vertical amplifier, when it is desired to synchronize the sweep with the signal; it may issue from whatever portion of the oscilloscope affords the 60-cps supply (usually a filament transformer in the power-supply section) when it is desired to synchronize the sweep with the 60-cps or line supply; or it may issue from any external source elected by the user. It is, therefore, evident that the operation of the selector switch is independent of the variable control; the former precedes the latter, for it is necessary to first determine the origin of the synchronizing voltage before adjusting its value for proper operation of the sweep system.

Another point of interest relative to the synchronizing selector switch is that it must be manipulated in conjunction with the switch that determines the signal input into the horizontal amplifier. If the latter is set so that it feeds the sweep voltage to the horizontal-deflection plates for the examination of a signal which is supplied to the vertical-deflection plates via the vertical amplifier, the synchronizing voltage should be secured from the vertical amplifier, that is, internally. Therefore the switch which determines the input to the synchronizing system must be set to INT. The remaining positions of this switch are similarly determined by the setting of the horizontal-amplifier input selector. We might also comment that, in a few instances, the synchronizing voltage selector switch is a part of the horizontal-amplifier input selector switch. Finally, in place of a rotary switch, the selection of the synchronizing signal source may be made by means of a three-position toggle switch.

The names assigned to these controls vary quite a good deal. Below is the listing of the names most commonly found on commercial oscilloscopes.

SYNCHRONIZING SYSTEM CONTROL IDENTIFICATIONS

FUNCTION	NAMES ASSIGNED
Select Synchronizing Signal Source	Synchronizing Sync Selector Sync Signal Selector H Sync/Sweep Selector Horizontal
Varies Synchronizing Signal Amplitude	Sweep Sync Sync Sync Adjust Sync Signal Amplitude Sync Lock Sync Signal Locking Sync Amplitude

Intensity-Modulation System Control Nomenclature

From the viewpoint of controls, the intensity-modulation system, even when it is identified as a *Z*-axis amplifier, is relatively simple. Having explained the basis of the action earlier in this chapter, it need not be repeated here, although we can profitably illustrate the circuit in elementary form. Two systems need be considered, although one of these is really so simple that it hardly justifies being identified as a system.

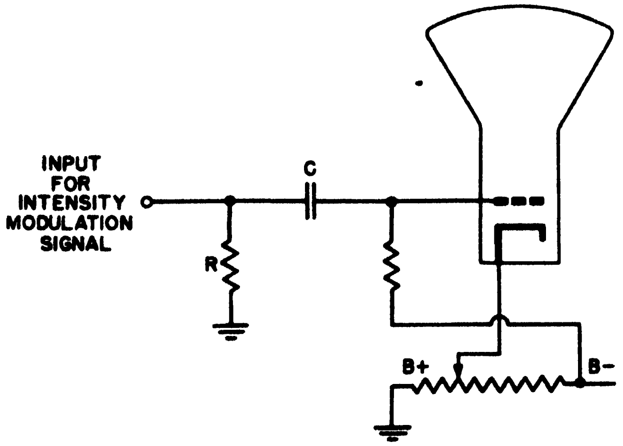


Fig. 10-27.—Simple intensity-modulation circuit.

The simpler of the two versions is shown in Fig. 10-27. It is an *R-C* network as shown, and serves as the means of injecting a signal into the control-grid circuit of the cathode-ray tube. The means of feeding the signal into the oscilloscope is via one or two terminals, usually only one. This terminal is labeled BEAM MOD or INTENSITY MOD, there being little choice between the two. This terminal is connected to what would be the "high" side of the modulation voltage supply circuit; the other leg of the circuit is joined to any of the ground terminals on the face of the panel. Sometimes the intensity-modulation binding post is located on the front of the oscilloscope and sometimes it is accessible at the rear, the former being the more common arrangement.

The second arrangement used for intensity modulation involves a simple amplifier between the input terminal and the circuit of the cathode-ray tube. This is the *Z*-axis amplifier. This leads to a change in the labeling at the input terminal. Instead of the markings previously mentioned, the identification may be Z AXIS AMPLIFIER. Like the simpler system, the input terminal usually is located on the panel, but occasionally it is mounted on a block accessible through the rear of the instrument housing.

The action of the intensity-modulation system is independent of the remaining portions of the oscilloscope. Therefore, control of the circuit is not vested in any switches which are tied into the other systems. Sometimes, however, a control marked BLANKING may be found on the panel. This is in the form of a simple switch which either completes the circuit or opens the circuit between the sawtooth oscillator and the grid-cathode circuit of the cathode-ray tube. When the circuit is closed, the negative swing voltage in the time-base oscillator (discharge period) is used as a negative pulse to extinguish the electron beam during the retrace time. When the switch is open, the retrace appears on the screen.

Power-Supply Control Nomenclature

Panel controls for the power-supply section of the oscilloscope are usually conspicuous by their absence. Some means for applying the operating voltages must exist and it does; it is an on-off switch which sometimes is individual and is located on the panel in the form of a toggle switch. In many cases, however, it is not visible as an individual switch because the function is made part of one of the other controls. In most instances ,it is a part of the control marked INTENSITY. The switch is mounted on the control and is activated when the control arm is advanced either to the extreme left position, which sets the switch to OFF, or if the control is advanced from its extreme left-hand position, when the switch is turned on. This is indicated by either an audible click, or a change in the tension of the control arm.

The labeling of the switch position is "OFF" when it is a part of another control. When it is individual, it bears the words ON-OFF. Sometimes, the power switch is labeled POWER and ON-OFF, the selection of the labeling being, as we have frequently stated, a matter of the individual designer's preference.

Control of the power-supply system is often vested in an automatic arrangement which is not visible on the panel. It is a safety measure and consists of a pressure-actuated switch which is located inside the cabinet, being mounted in such position on the chassis that the power is turned on only when the outside housing, or the cabinet, is properly in place. This is a safeguard against possible shock in the event that the cabinet may have been removed and an attempt made to handle the components.

FUNCTIONAL ORGANIZATION OF THE BASIC OSCILLOSCOPE

Let us now examine the possible arrangements of the different sections shown in Fig. 10-1 and see what systems can be derived to perform varied functions. The device contains all of the essential sections necessary for very many general applications. It is a general-purpose oscilloscope possessing performance capabilities which are established mainly by the performance specifications of three major sections, the vertical amplifier, the horizontal amplifier, and the time-base oscillator. The remaining facilities stem from the switching means, as do those capabilities which involve the sections we have mentioned. No matter what is the intended application, it requires the manipulation of the switches and control in order to organize functionally the device to suit the need. These are the details we shall discuss in this section.

In order to show the manipulation of switches and controls, especially the former, we must imagine a certain specific oscilloscope. Obviously it is impossible to deal with each specific device, for, as we have shown earlier, the labeling may differ, and this would make identification almost if not entirely impossible. The same is true of the variables of design as encountered in the many devices, and the differences in the circuits controlled by any one switch. In other words, it leads us to only one means of presenting the possible functional organizations, namely, the arbitrary selection of a device which can be used as the sample. This is the device we have depicted in block form in Fig. 10-1 and which we repeat here as Fig. 10-28 simply as a matter of convenience.

It is true that this figure does not afford every variety of system structure but it is a fairly good compromise. It contains enough to enable us to show very many possible combinations of systems, certainly enough to create familiarity with the basic elements of the cathode-ray oscilloscope. Subsequent discussions in this chapter will cover the modifications of the basic instrument which remove it from one category and place it within another.

Earlier in this chapter we placed the dot of light on the screen of the cathode-ray tube. All future illustrations of the screen with the beam deflected or undeflected will be predicated on the assumption that the operating controls have been set so that the electron beam is properly focused, the spot on the screen has adequate brilliance, and its undeflected location is at the center of the screen.

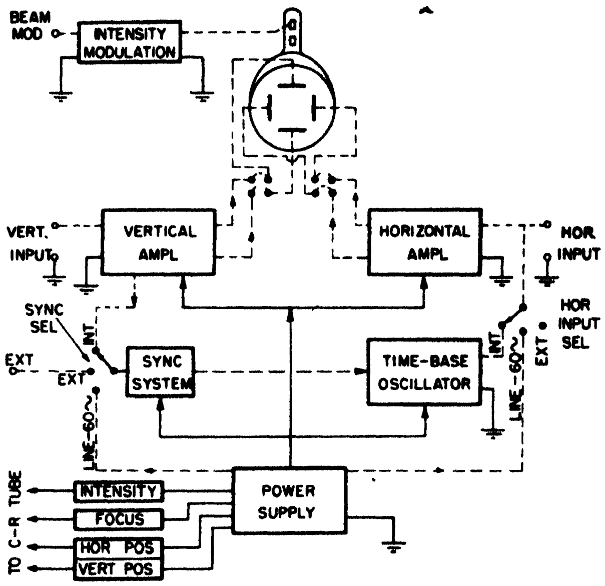


Fig. 10-28.—Block diagram of the basic oscilloscope.

Direct Connection to Deflection Plates

The simplest organization of the sections in the basic oscilloscope is the cathode-ray tube. This is shown in Fig. 10-29A and B. When so employed, it is an indicator, although the phenomena which it may display are numerous. As to possible reasons for using the tube

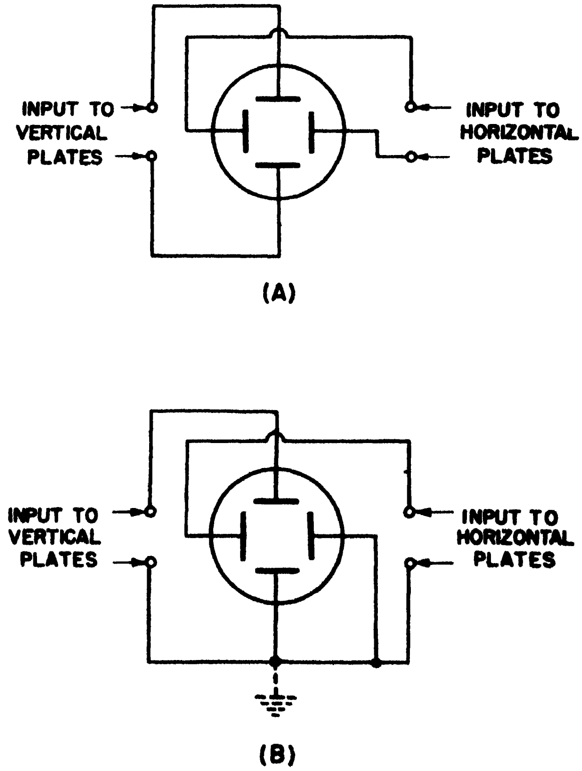


Fig. 10-29.—Inputs to the deflection plates, (A) balanced, (B) unbalanced.

alone, the following are some of them. Perhaps the cathode-ray tube in one device is used as a replacement for a defective tube in another, for, as can be seen, no restrictions are indicated relative to the source of the signals which are applied to the vertical- and horizontal-deflection plates. These signals may originate in the vertical and horizontal amplifiers of another oscilloscope, or in a device completely foreign to an oscilloscope, as, for example, a transmitter.

Another reason for using the cathode-ray tube in this manner is that the frequency of the signals to be displayed exceeds the response capabilities of the vertical and horizontal amplifiers. A typical example of this is the development of the trapezoidal pattern showing the modulated output of an amplitude-modulated transmitter. How this application is accomplished is discussed completely in Chapter 18, but several typical traces of this kind are shown in Fig. 10-30. Their meanings will not be treated here because they receive attention later.

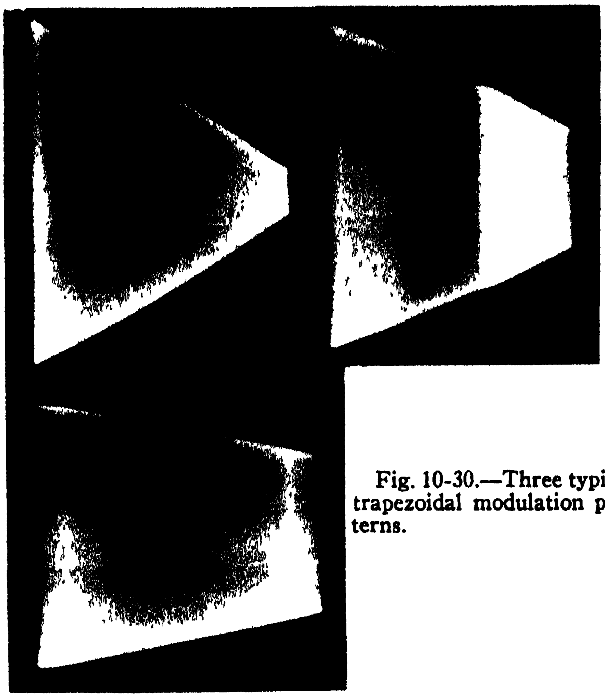


Fig. 10-30.—Three typical trapezoidal modulation patterns.

The illustration of the deflection system circuits does not indicate any means for controlling the levels of the signals applied to the vertical- and horizontal-deflection plates; it is assumed that the signals are of proper level, or that the source contains the level controls. Thus if the patterns in Fig. 10-30 represent the carrier signal fed into the vertical-deflection plates and the corresponding modulating signals fed to the horizontal-deflection plates, the levels of the respective signals are arranged so that at their origin the final trace has the correct dimensions on the screen.

The deflection elements are shown connected in two ways. Fig. 10-29A typifies the modern version, wherein utmost flexibility is offered. Such an arrangement enables the use of either balanced or unbalanced input systems. This arrangement permits the application of the elements to either grounded or ungrounded systems.

The older version of deflection circuits is shown in Fig. 10-29B. Two plates, one from each deflection pair, are joined and connected to ground. Consequently, only one plate of each system is above ground and the arrangement is called unbalanced. Although relative to the tube display, both arrangements generally behave alike, the circuit of Fig. 10-29B introduces limitations. Since the two deflection systems are internally interconnected, a ground on one circuit will automatically ground the other circuit. This does not matter in some applications, but in others it may introduce complications because the system which feeds the voltages to the vertical- and/or horizontal-deflection plates may not permit grounding. Due to the ground at the common junction, even though individual terminals are usually provided for each pair of plates, the deflection system is restricted to connection between *ground and some other point,* rather than between any two points, as is permitted by the system shown in Fig. 10-29A.

In Fig 10-29B, the junction between the two plates is assumed to exist inside the tube. Inasmuch as we are speaking about the deflection system without recourse to any amplifiers, any form of circuit wiring in the oscilloscope proper which ends up in an electrical system corresponding to Fig. 10-29B, but in which the actual wiring of the deflection plates inside the tube is such that they terminate in individual connection terminals, such a tube would be symbolized by Fig. 10-29A rather than by Fig. 10-29B. The reason for this is the fact that when the deflection plates are used apart from the remainder of the oscilloscope, contact with the plates is made at the individual terminals provided for that purpose.

The deflection-plate connection terminals are often located on terminal boards accessible through the sides or rear of the oscilloscope cabinet. Sometimes they are available on the front panel and still other times they are pin-jacks accessible through holes in the sides of the cabinet. When the terminals are located on a terminal board, the jumpers which join the deflection plates to the amplifier leads must be removed in order to isolate the deflection system from the amplifier. Occasionally this operation is obviated by the presence of two se-

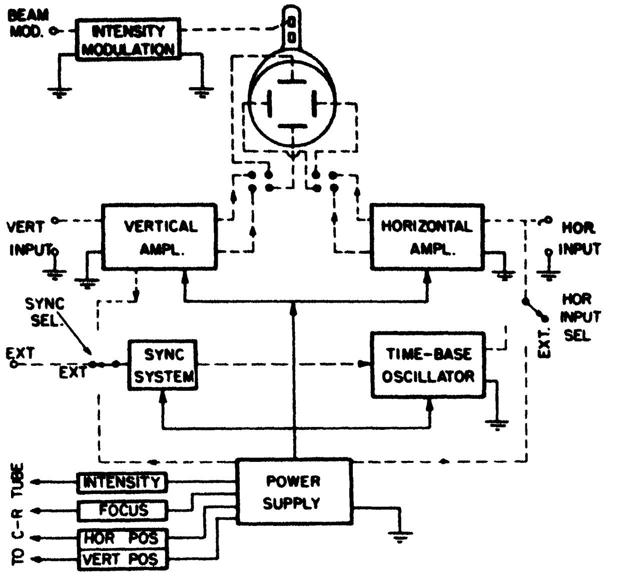


Fig. 10-31.—Block diagram of the basic oscilloscope arranged for direct connection to the deflection plates.

lector switches on the front panel, by means of which isolation of the deflection plates and the amplifier is accomplished.

As to the setting of the switches on our hypothetical basic oscilloscope, the block diagram shown in Fig. 10-31 may be used for reference. Only the active position of the respective switches is shown, assuming that the switches indicated will be set to these positions. In this specific case, it is simply a matter of convenience to set them as shown, since the cardinal operation is the removal of the jumpers from the terminal boards. A comparison of Fig. 10-31 with the basic illustration of Fig. 10-28 will illustrate just what we mean.

When signals are fed directly to the deflection plates without using the amplifiers, the problem of signal amplitude arises. In order that the cathode-ray tube be usable in this fashion, certain minimum signal requirements exist. First, the signal level must be sufficient to give a trace at least 1 inch high and 1 inch wide. This is the *absolute minimum* and is most certainly *not a convenient size for study purposes;* it is entirely too small. Of course, the dimensions of the screen have a great bearing on what is a suitable size; the greater the screen dimensions, the larger the image should be. For all general purposes, a trace that is 2 to 3 inches high and of the same width will be satisfactory. We might suggest that a satisfactory image size is approximately 60 to 70 per cent of the tube-face area.

Direct connection can be used only when the signal voltages applied to the tube amount to at least the deflection-factor rating corresponding to a 1-inch deflection. Since deflection factor is based on applied d-c voltage, a deflection of 1 inch can be attained when the *peak-to-peak* a-c voltage equals the required d-c voltage. Thus if the rating is 100 volts d.c. per inch, the same over-all dimensioned trace will be obtained with an a-c voltage 100 volts peak to peak, or an rms value of approximately 35.7 volts. This magnitude of voltage is the reasonable *minimum* and should not be erroneously construed as the maximum.

The Cathode-Ray Tube With Amplifiers

A possible organization of the basic oscilloscope is shown in Fig. 10-32. Here, because the frequency of the quantity to be examined exceeds the capabilities of the vertical amplifier, the signal is fed directly to the vertical-deflection plates. It is possible that the signal to be examined is of such level that it cannot be attenuated properly through the vertical amplifier, or the conditions of operation are such that the vertical-amplifier input system specifications do not allow it to be connected across the source of the signal. Whatever the cause may be, the signal is fed directly into the vertical-deflection system.

On the other hand, the horizontal amplifier is used, although the time-base oscillator is not. Its frequency range may be inadequate or the device may be defective, so that an external time-base oscillator is used. Some of these applications are not very likely, but since we are dealing with the subject from the broadest viewpoint, we must illustrate the greatest number of sectional organizations possible within the basic oscilloscope.

It is not imperative that the signal applied to the horizontal amplifier be a time-base voltage secured from an external source; it can be any other signal which is to cause the horizontal displacement of the beam. Finally, the signal being fed into the vertical-

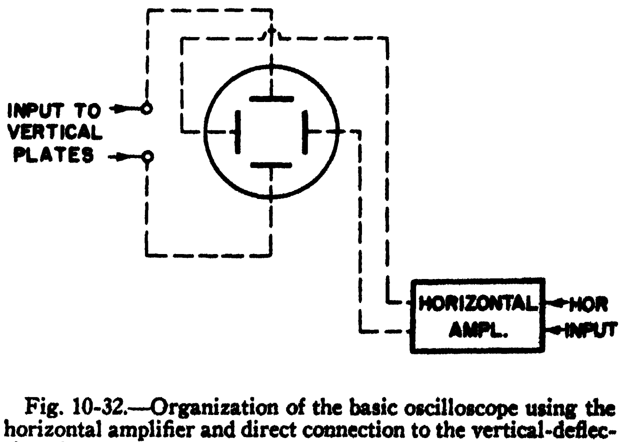


Fig. 10-32.—Organization of the basic oscilloscope using the horizontal amplifier and direct connection to the vertical-deflection plates.

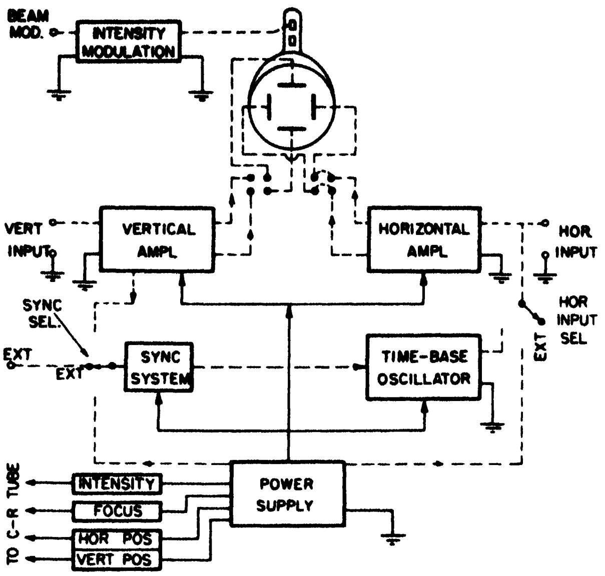


Fig. 10-33.—Block diagram of the basic oscilloscope corresponding to Fig. 10-32.

deflection plates may issue from the vertical amplifier in some other oscilloscope. In other words, a combination of two devices are being utilized for a special purpose.

Among the capabilities of the horizontal amplifier is the control of the level of the signal being fed into the cathode-ray tube via that channel. This is not indicated because the gain control is inherent to the amplifier, and it is understood that after the switches have been properly set, as shown in Fig. 10-33, the gain control will be used as required. An example of a pattern developed under such conditions, as, for example, the wave envelope of a high-frequency modulated wave, is shown in Fig. 10-34. It is assumed that the level of the modulated carrier is sufficient to permit direct application to the vertical-deflection plates; the time-base voltage is secured from an external source and it is assumed that it has been set to the correct frequency relative to the frequency of the modulating voltage. This relationship, rather than the carrier frequency versus the sweep frequency, determines the number of cycles of the modulated wave which appear on the screen. The use of the system shown is unlikely, but since we are speaking about possibilities in system organization, it is shown.

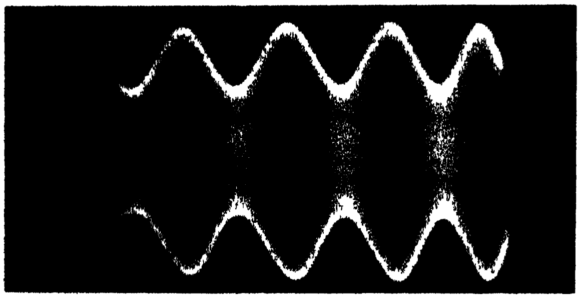


Fig. 10-34.—Wave envelope of a high-frequency modulated wave.

Vertical Deflection Only

Still another arrangement of the basic oscilloscope is shown in Fig. 10-35, and the corresponding switch positions are shown in Fig. 10-36. Here only the vertical amplifier and the vertical-deflection plates see service, and consequently the deflection will be in one direction only, vertical. Since the vertical amplifier is in the circuit, it determines both the upper- and the lower-frequency limits of the signal voltage applied to the input. When a signal is applied to the input of the vertical amplifier, the trace on the screen takes the form of a single straight line running vertically across the screen. Its length is determined by the amount of signal voltage applied to the deflection plates, but since

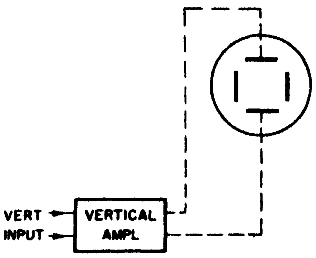


Fig. 10-35.—Organization of the basic oscilloscope using the vertical amplifier and deflection plates only.

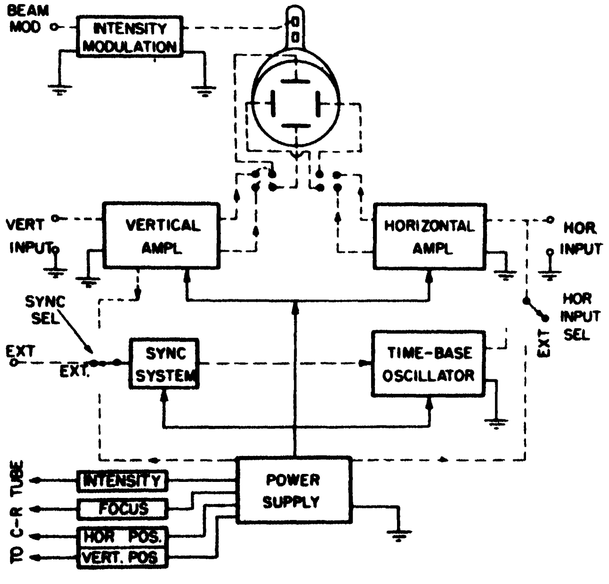


Fig. 10-36.—Block diagram of the basic oscilloscope corresponding to Fig. 10-35.

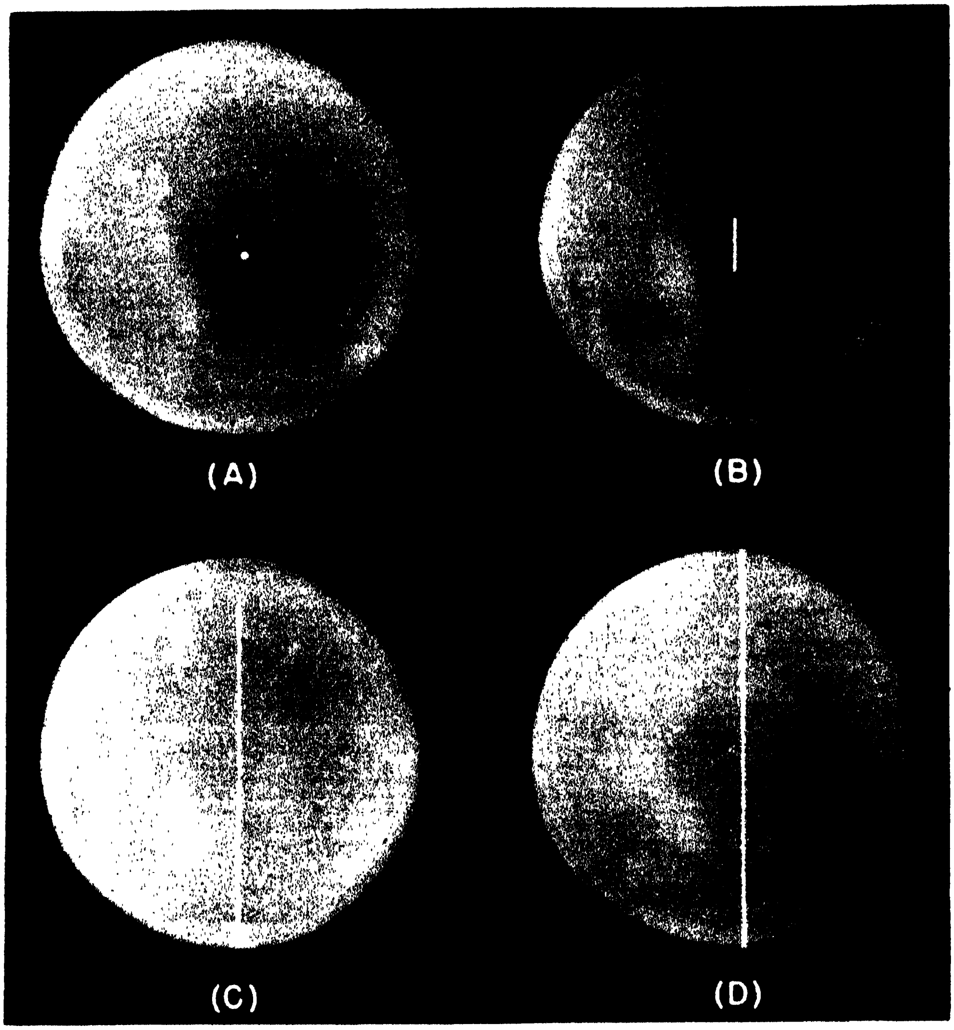


Fig. 10-37. — Oscillograms taken with organization as in Fig. 10-35; (A) zero setting of the vertical gain control, (B) insufficient height of the trace, (C) satisfactory height of the trace, and (D) excessive height of the trace.

the same signal also exists at the input of the vertical amplifier (across the vertical-amplifier gain control), where it may be more or less than the amount of voltage at the deflection plates, it stands to reason that, in the final analysis, the setting of the vertical gain control will determine the dimension of the trace.

This might be obvious from all that has been said before, but it warrants mention and illustration just the same. Oscillograms (A), (B), (C), and (D) in Fig. 10-37 indicate the appearance of an a-c voltage without time-base deflection for several settings of the vertical gain control. These are: zero setting of the vertical-gain control (A), insufficient height of the trace (B), satisfactory height of the trace (C), and excessive height of the trace (D). In (D) the trace spreads vertically beyond the limits of the tube screen.

The previous reference to the normal height of the trace means a length which is about 2 inches high on a 3-inch screen, and 3.5 to 4 inches high on a 5-inch screen; in general, about 60 to 70 per cent of the screen diameter.

We might mention that if the frequency of the voltage applied to the vertical-deflection plates is very low, say 2 or 3 cycles or less, the trace will not be a fixed line; it will "breathe," which is what we call the action of expansion and contraction of the line. No matter where the vertical gain control may be set, it will still contract and expand at a rate corresponding to its frequency. The ability to see the action of amplitude changes is due to the low frequency. At high frequencies, the change takes place too rapidly, and a complete line appears on the screen.

In view of the contents of Chapter 19, it is of interest at this time to comment upon the use of the sectional organization shown in Fig. 10-35 for the measurement of the peak-to-peak values of signal voltage. This particular application is not an important one because many similar measurements can be made more conveniently with simpler equipment, certainly with equipment which will not load down the source of the voltage as much as a cathode-ray oscilloscope might. We are referring to the high-impedance vacuum-tube voltmeter. However, it is becoming common practice in the more elaborate oscilloscopes to provide for signal-level measurements as a convenience while making other determinations, thus obviating the necessity for using supplementary equipment. This is especially significant since such measurements may be made while observing the waveform of the signal being investigated, or any portion of the signal.

The above is not evident in Fig. 10-37 since the time-base dimension is absent, but the idea can be gathered from the statement that since the full length of the trace represents the peak-to-peak value, half the trace height is the peak value of one alternation, and if the length of the trace each side of the undeflected position of the spot is converted into voltage, any fraction of that dimension can likewise be converted into voltage on a proportional basis. Devices for just such measurements are available as supplementary apparatus for oscillographic use. In fact, they are to be found as parts of complete oscilloscopes. (They are known as calibrators. See Chapters 14 and 19.)

Frequency Comparison

A very commonplace arrangement of the sections of an oscilloscope is shown in Fig. 10-38. The vertical and horizontal amplifiers are present but no time-base source. This arrangement is identified at the head of this paragraph as being suitable for frequency measurements or frequency comparison. This is true, but it also is suitable for any other operation which demands the use of only the vertical and horizontal amplifiers from among the units within the device. Visual alignment is one of these when the horizontal sweep signal is supplied by the sweep generator. The switch positions are shown in Fig. 10-39.

Obviously the frequencies applied to the vertical and horizontal deflections must be within the response of the two amplifying systems. While it is customary to think about sine waves when making frequency comparisons, nonsinusoidal waves may be used.

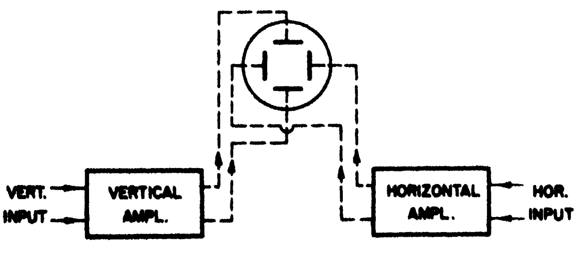


Fig. 10-38.—Organization of the basic oscilloscope using both horizontal and vertical amplifiers.

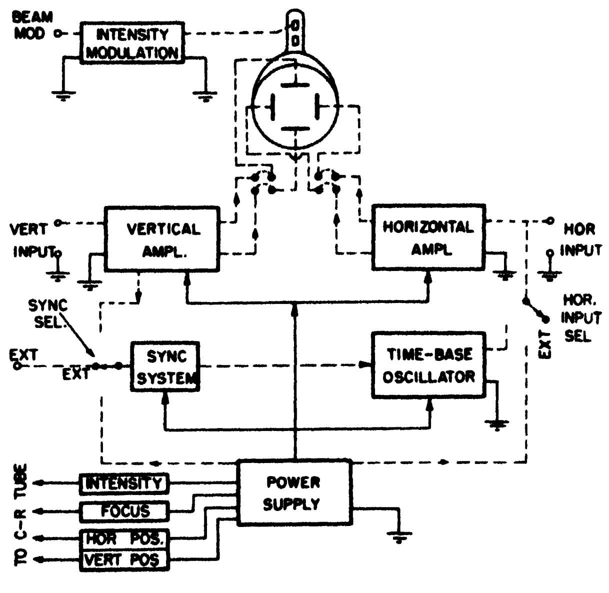


Fig 10-39.—Block diagram of the basic oscilloscope corresponding to Fig. 10-38.

Chapter 12 is an elaborate presentation of applications which would use the oscilloscope organization illustrated in Fig. 10-38 and that chapter also contains very many illustrations depicting phase and frequency measurements. In order that we remain consistent with what has gone before, Fig. 10-40 shows six patterns resulting from the combination of different types of

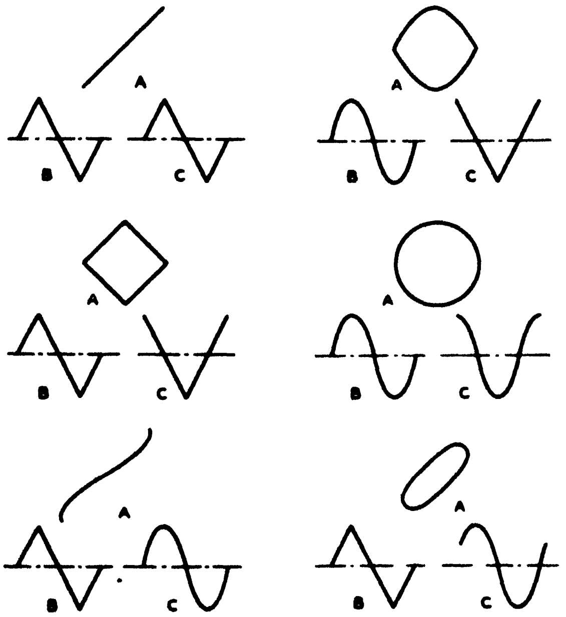


Fig. 10-40.—Six oscillograms taken using the arrangement of Fig. 10-38; in each case *A* is the resultant, *B* is the shape and relative amplitude of the wave fed into the vertical amplifier, and *C* is the wave fed into the horizontal amplifier.

waves fed to the vertical and horizontal amplifiers organized as indicated. In each case *A* is the resultant, *B* is the shape and relative amplitude of the wave fed into the vertical amplifier, and *C* is the wave fed into the horizontal amplifier. The explanations for these patterns and many more, simple as well as complex, are given in Chapter 12.

Horizontal Deflection Only

Two possible organizations of the sections of the basic oscilloscope are shown in Fig. 10-41A and 10-41B. These have little value in oscillography, other than as a means of familiarization with the operation of horizontal deflection due to the sawtooth time-base oscillator or the sine-wave time-base voltages. It is shown specifically for that purpose. Examination of the block diagrams discloses that the vertical-deflection plates receive no voltage, whereas the horizontal-deflection plates may receive a voltage from either the sawtooth time-base oscillator or from the sine-wave time-base source through the horizontal amplifier. The choice of either kind of signal input to the horizontal amplifier is made by means of the horizontal amplifier input selector switch as shown in Fig. 10-42A and B, respectively.

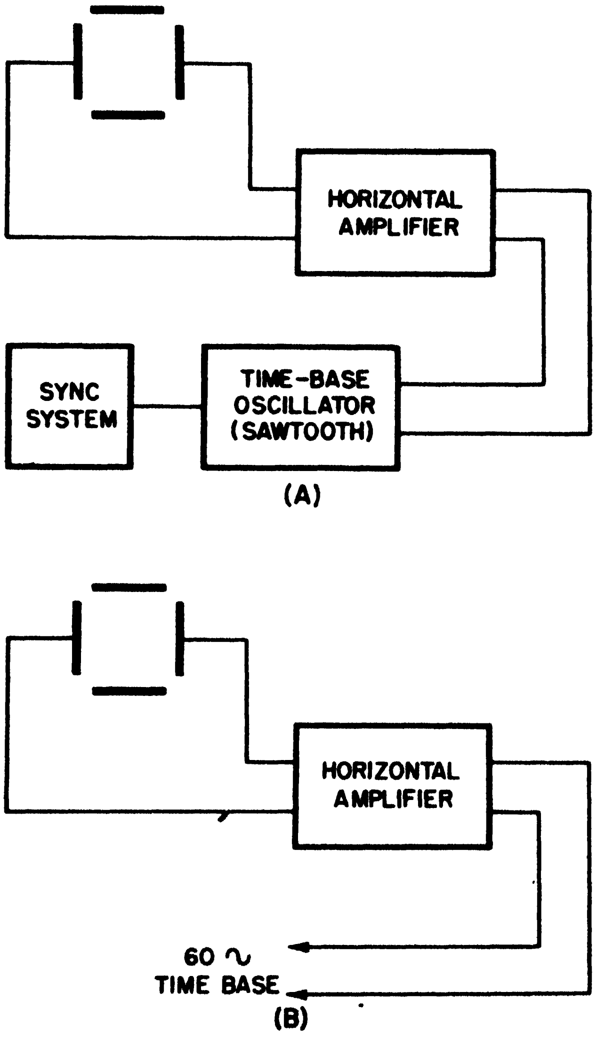


Fig. 10-41.—Possible organizations of the basic oscilloscope; in (A) the blocks are arranged for internal synchronization, and in (B) they are arranged for 60-cps synchronization.

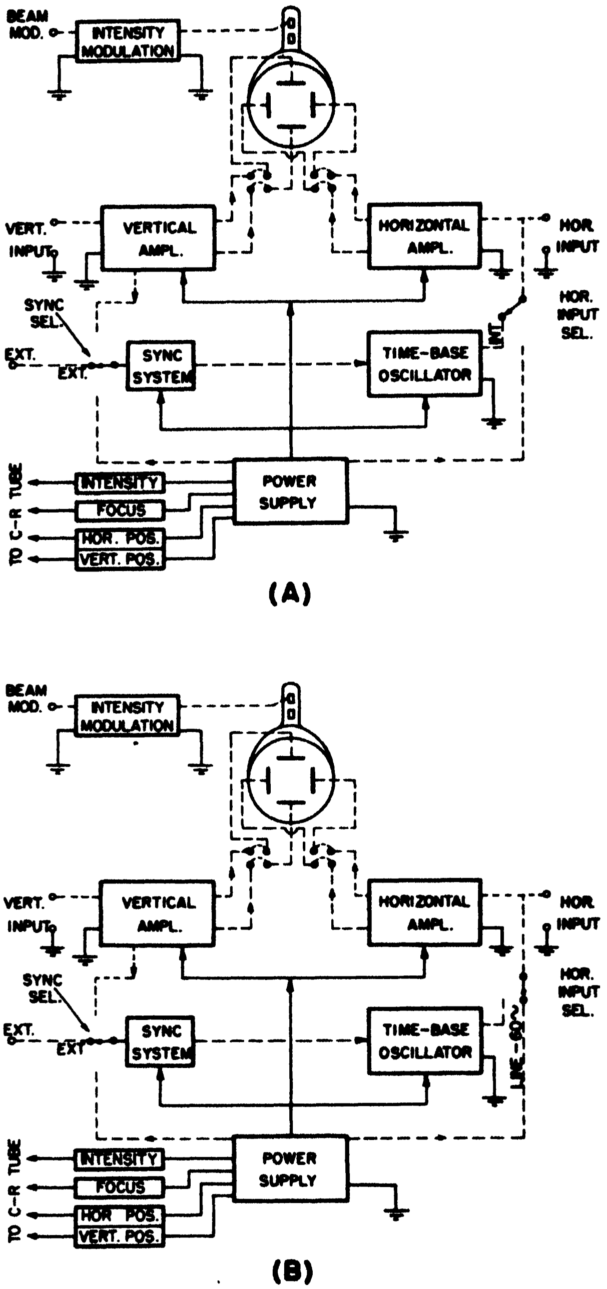


Fig. 10-42.—Block diagrams of the basic oscilloscope corresponding, respectively, to the organizations of Fig. 10-41.

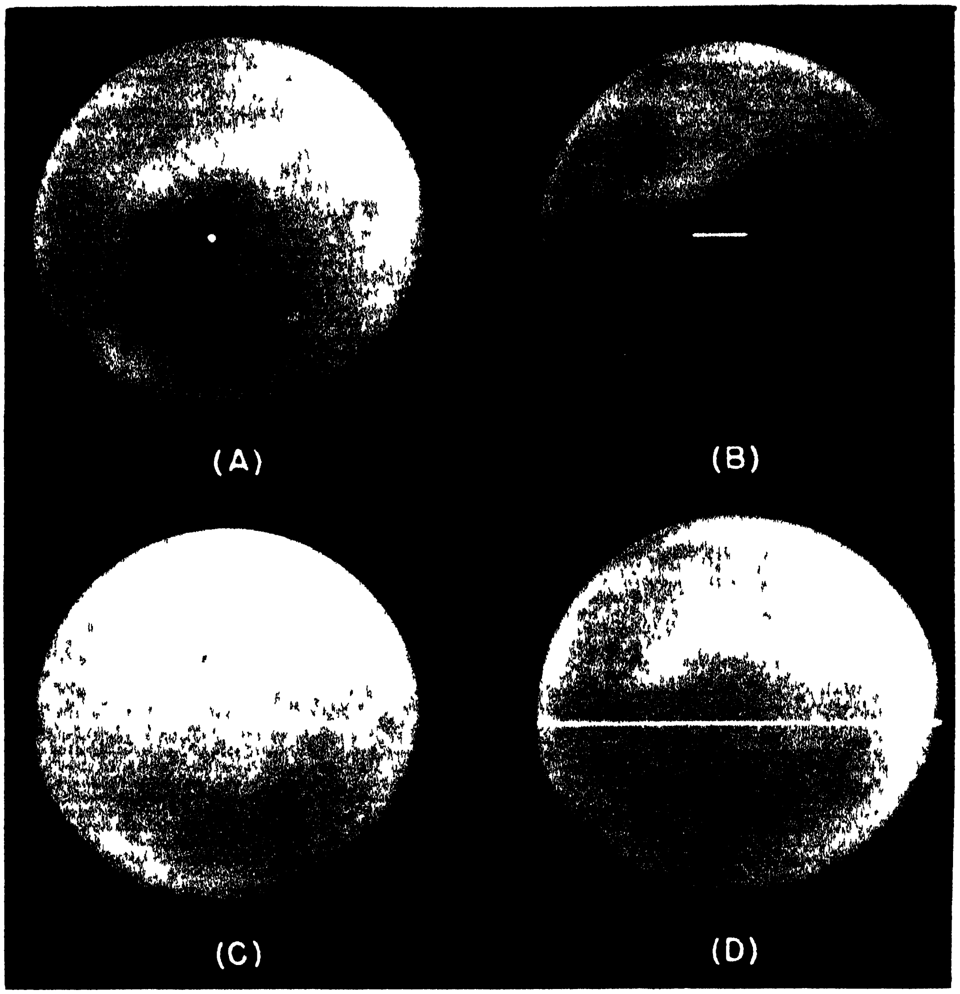


Fig. 10-43.—Oscillograms demonstrating various conditions of horizontal gain without any vertical input; (A) shows zero setting of the gain control, (B) shows insufficient amplitude, (C) is the correct amplitude, and (D) shows excessive horizontal amplitude.

In the absence of the vertical deflection, the signal voltage applied to the horizontal-deflection plates appears on the screen as a horizontal trace. As in the case of the vertical amplifier (Fig 10-35), the magnitude of this trace is determined by the amount of signal voltage applied to the horizontal-deflection plates, and, again as in the case of the vertical-deflection plates, the horizontal-deflection voltages appear originally at the input of the horizontal amplifier. Consequently, the setting of the horizontal-amplifier gain control determines the magnitude of horizontal deflection and can vary it from zero to the value required to spread it beyond the limits of the screen. This is shown in Fig. 10-43 as (A), (B), (C), and (D) where (A) corresponds to the zero setting of the horizontal gain control, (B) is insufficient amplitude, (C) is correct amplitude (60 to 70 per cent of the screen diameter), and (D) is excessive horizontal amplitude because the horizontal gain control is advanced too far and the trace extends beyond the limits of the screen.

Which of the time-base sources is being used for the development of the traces in Fig. 10-43 is immaterial, since both will look alike. While the character of these voltages is different, one cannot be distinguished from the other because the vertical displacement necessary to show amplitude variations is lacking. All kinds of a-c voltage would look alike under such circumstances, appearing as a single line trace with a length determined by the value of signal voltage fed into the horizontal gain control, the setting of this control, and the gain of the amplifier. The exception to the appearance of the trace is the very-low-frequency sawtooth voltage. If the frequency is one or two cycles, the trace will be a dot which will be in motion, but if the frequency is higher, the trace will be a complete line. Varying the horizontal gain control when the trace is a discrete dot of light advancing across the screen will determine the length of its path.

WAVEFORM PRESENTATION

The organization of the sections of the basic oscilloscope for the examination of waveforms, or for the examination of actions which have been converted into

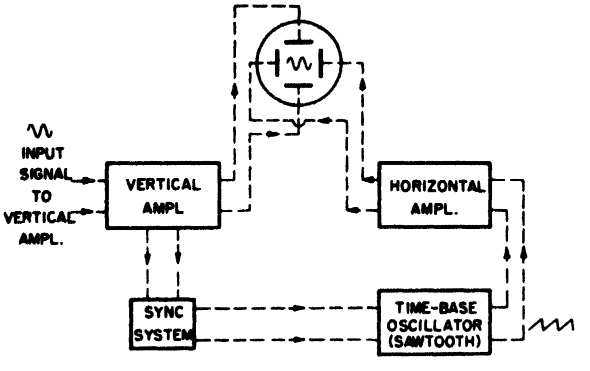


Fig. 10-44.—Organization of the basic oscilloscope for the display of waveforms.

equivalent voltages, is shown in Fig. 10-44. This is the most prominent application of the oscilloscope. For such determinations, the cathode-ray oscilloscope is beyond comparison with any other instrument. However, since we are dealing with a basic example of the device, we are forced to restrict the extent of the application to those functions which are within the capabilities of this category of instruments. This simply means that we shall not deal with the display of nonrecurrent, short-duration pulses which require triggered sweeps. Instead, we shall devote our attention to those quantities that can be displayed by the conventional repetitive sweep, the sawtooth kind first, then the sine-wave variety. The more elaborate versions will be treated separately later in this chapter, when we analyze the manner in which the basic oscilloscope is modified so as to become a more elaborate instrument.

The significant additions to the block diagrams shown so far are the time-base oscillator and the synchronizing system. These two are required if the voltage applied to the vertical amplifier is to be shown relative to time. The switch positions for examination of a wave using the sawtooth sweep are shown in Fig. 10-45.

In order that we advance in a logical manner, it will be necessary at this time to refer back to Chapter 9 for just a brief moment. That chapter explained the manner in which the time-base voltage deflection, acting in conjunction with the deflection due to the vertical voltage, resulted in a trace which depicted amplitude variations versus time. We illustrate this in Fig. 10-44 by a sine-wave voltage at the input of the vertical amplifier, a sawtooth voltage at the output of the time-base oscillator, and the final sine-wave picture on the screen of the tube.

At this point, let us digress for a moment in the chain of events and make a few comments concerning the relationship between the frequency of the signal under observation and the frequency setting of the time-base oscillator. As a beginning, let us recall that the usual arrangement for the selection of the frequency of the latter is in the form of two variable controls. The multiposition switch control selects the frequency range; the continuously variable control affords "fine" tuning of the oscillator system so that the precise frequency necessary for displaying a clear, easily readable trace is attained. The frequency of the time base relative to the frequency of the signal fed into the vertical amplifier determines how many cycles of the quantity being examined will appear on the screen.

As a general rule, the frequency calibrations of the time-base oscillators are rough; they approximate the panel markings but cannot be preset with accuracy for calibration purposes. This does not indicate faulty design, for, after all, the device may be a free-running multivibrator until it is locked at a specific frequency by the synchronizing signal. This signal tends to vary the frequency of the sweep oscillator, and, therefore, prevents any attempt to use the frequency markings on the panel for precise calibration or frequency determination. At best, these units bear tolerance limits and this too tends to aggravate the situation. We are speaking about the general-purpose devices typified by the basic oscilloscope, although we can go so far as to say that only a special category of devices affords sweep frequencies which approach a high order of accuracy. In the final analysis, great accuracy in calibration of the general-purpose sweep-voltage source is not essential because each application demands individual adjustments so as to place the desired trace on the screen.

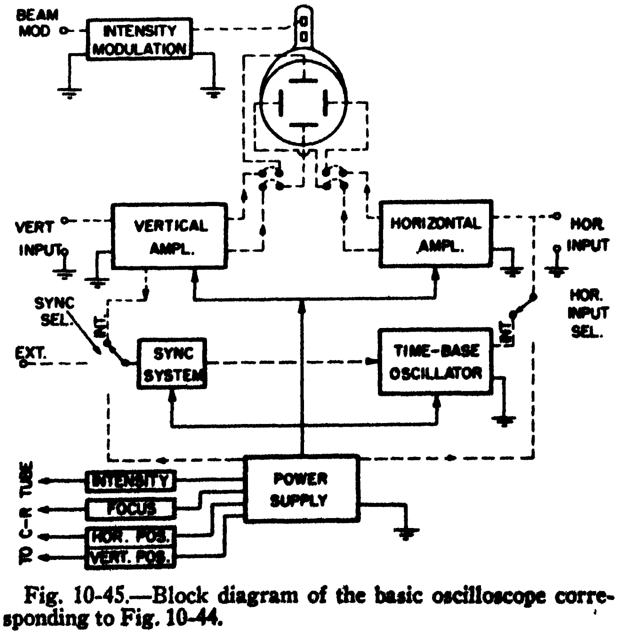


Fig. 10-45.—Block diagram of the basic oscilloscope corresponding to Fig. 10-44.

Such adjustments must, however, be correct and are attained best by careful attention to the frequency setting of the sweep oscillator with minimum synchronizing voltage. The adverse effects of excessive synchronizing voltage can be as great as insufficient synchronization.

Relationship Between Sweep and Signal Frequencies

Speaking in generalities, with more elaborate explanations to follow, the relationship between the frequency of the signal being applied to the vertical-deflection plates and the frequency of the time base determines the number of cycles of the signal which will appear on the screen. In *all cases* of waveform display, the frequency of the sweep voltage is either less than that of the signal, or equal to that of the signal, but never higher than that of the signal. If the sweep frequency is higher than the signal frequency, only portions of the signal cycle will appear on the screen and the interpretation is extremely difficult.

As to what should be the correct frequency status between these two signals, it is a matter of individual choice, although certain conditions inherent in oscilloscopes dictate a certain treatment. This is the appearance of at least 2 cycles of the signal wave and preferably 3. The reasons for this will be shown soon.

The suggestion to display 3 cycles of the signal being examined is not a rigid rule; it can be more or less, to suit the viewer. In fact, sometimes the design of the time-base oscillator prevents this. For example, if the frequency of the signal to be examined is 10 cycles, the sawtooth oscillator must have a low limit of 3 cycles in order to display 3 cycles on the screen, which, in turn, means that the horizontal amplifier must pass a 3-cycle fundamental sawtooth voltage, and the vertical amplifier must pass a 10-cycle signal. The same controlling factors would apply for whatever may be the signal frequency. At the high end, the usual controlling factor is the upper-frequency limit of the sweep oscillator. If a 600-kc wave is to be examined with a sawtooth sweep rated at a top frequency of 30 kc, 20 cycles of the signal will be the fewest capable of display. This is too many, even if it is possible to spread the horizontal deflection so that only a part of the entire display appears on the screen.

Several examples of the manner in which the ratio between the time-base oscillator and signal frequencies determines the number of cycles that appear on the screen might aid comprehension. Consider the following:

Sweep Frequency (cycles)	*Signal Frequency (cycles)*	*Number of Cycles on screen*
1	1	1
1	5	5
3	9	3
11	22	2
150	300	2
2,400	7,200	3
10,000	30,000	3
10,000	100,000	10

These are random examples and, in every case, the frequency of the sweep is below the frequency of the signal. Pictorially, some of these appear in (A), (B), (C), and (D) of Fig. 10-46. The greater the ratio between the signal and the sweep frequency, the greater will be the number of cycles on the screen. This figure shows a sample of a single-cycle trace (A), multiple-cycle traces (B) and (C), and in (D) a trace which is nothing more than a rectangle of light because the individual cycles blend into each other due to the extremely high ratio. From all of this it is possible, generally, to conclude that a trace with more than 5 or 6 cycles contains too many cycles.

Referring once more to the tabulation of frequencies, these are by no means the specific frequencies to which the time-base oscillator and the signal sources are limited. Since the time-base oscillator is continuously variable in frequency, its setting is determined by the constants arranged by the position of its controls. In like fashion the frequencies, which can be accepted by the vertical amplifier, cover a range and *any* frequency within this range may be applied to the input of the vertical system.

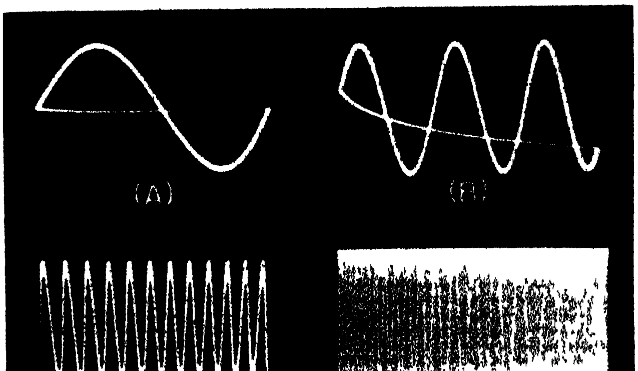


Fig. 10-46.—Illustrating various sweep-to-signal frequency ratios.

Integral and Fractional Sweep-Signal Frequency Ratios

Now, it is important to understand that some form of pattern will exist upon the screen for all ratios between the sweep frequency and the frequency of the wave to be examined. However, the nature of these patterns is too complicated to be of any value and, as a general rule, the patterns are moving. To establish the waveshape of the voltage applied to the vertical-deflection plates, it is necessary that the pattern appearing upon the screen be the simplest to show the waveshape. Such a pattern is established when the relation between the signal and sweep frequencies is a simple integral ratio, not a fractional ratio. When the former ratio exists, such as 1:1, 2:1, 3:1, 4:1, 5:1, etc., and the lower frequency is that of the sweep circuit, the maze of lines moving across the screen will develop into a simple waveform. Whether or not this wave will be sinusoidal or distorted depends upon a number of factors, which need not be discussed at this time. At any rate, the pattern desired is that which has a single line waveform of one or more cycles.

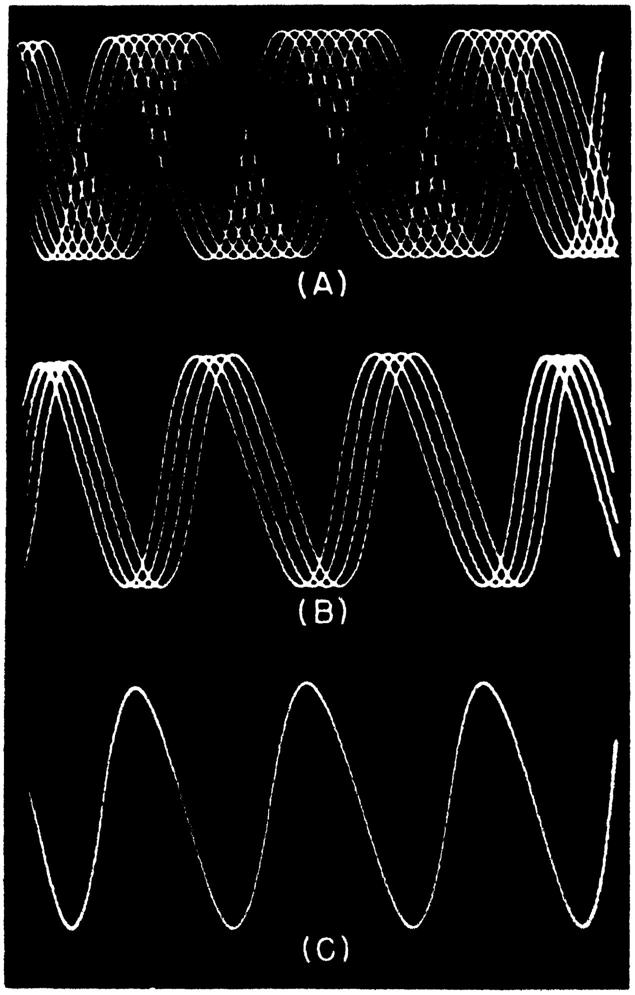


Fig. 10-47.—Three stages in the synchronization process.

The greater the ratio between the sweep frequency and the frequency of the voltage under observation, the greater is the number of lines which appear in the pattern during the time that the fractional ratio exists between the two frequencies. However, as the sweep frequency is changed and approaches that value which provides the simple integral ratio, the number of lines in the pattern becomes less and less, until a single line waveform pattern appears. This adjustment is reached by manipulating the fine frequency control. This is shown in three stages in Fig. 10-47.

Keeping the Pattern Stationary (The Synchronizing Control)

Using the time-base oscillator properly is very important to the fulfillment of the purpose of the oscilloscope. As individual as this section may seem in function, nevertheless, it is very closely associated with the synchronizing system or control. Proper selection of the time-base frequency relative to the signal frequency can result in a stationary pattern, but, due to unavoidable minor variations in operating voltages, the time-base oscillator will not remain constant in frequency. Instead it may require repeated readjustment, so that the pattern will remain stationary long enough for study.

The electrical conditions, created in the time-base oscillator when the synchronizing signal is injected, are explained thoroughly in Chapter 11, but in the meantime we can discuss the effects. One of these is constancy of frequency as long as the synchronizing signal is being applied; the other, and equally important, is that the synchronizing signal is capable of changing the frequency of the oscillator so that it will keep in step, if for some reason some variation in frequency would be occasioned by some other related change in operating conditions. A third and equally important characteristic is that the *amount of synchronizing voltage* injected into the multivibrator determines the behavior of that oscillator, and in that way the shape of the pattern.

Variables in Waveform Display

The content of the previous paragraph involves so many operating variables that this is an excellent point in the text to introduce a number of operating features which will demonstrate the application of the oscilloscope to waveform display, and, in general, describe the manner of handling the different controls located on the panel of the basic device. Just as a reminder, the time-base voltage is of sawtooth character until a change is stated.

Let us now assume the following set of conditions. The intensity, focus, and positioning controls are set so that the correct dot of light appears at the center of the screen. Then:

1. An audio signal source is connected to the input terminals of the vertical amplifier. The generator is placed into operation so that a signal of reasonable strength is available at its output terminals.
2. The vertical-amplifier gain control is advanced to about one-third its full range.
3. The HOR-INPUT selector switch is set at INT as indicated in Fig. 10-45.
4. The SYNC SELECTOR switch is set to INT as indicated in Fig. 10-45.
5. The SYNC control is set slightly beyond its 0 setting (or just in the + or − zones indicated) on the panel.
6. The rough frequency control of the sweep oscillator is set to that position which approximates 750 cps.
7. The horizontal-amplifier gain control is advanced just slightly beyond its minimum setting.
8. The fine frequency control of the sweep oscillator is then adjusted slowly until a single line pattern like that shown in Fig. 10-48A appears on the screen.

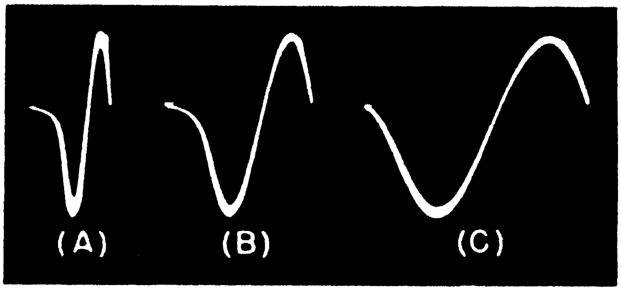


Fig. 10-48.—Illustrating the improvement in the pattern as a result of increasing the sweep amplitude.

9. A slight readjustment of the synchronizing control and the fine frequency control will result in a stationary pattern. (This is a single cycle of the signal applied to the vertical amplifier.)
10. A slight advance of the intensity control may be necessary in order to develop the satisfactory brightness of the trace since the motion of the electron beam reduces the time of screen excitation.

Obviously, the representation of the single cycle is not the most perfect-looking wave. There are several reasons why a symmetrical wave is not upon the screen. Although we do not know the characteristics of the signal available from the signal generator, we assume it to be a sine wave and the trace on the screen certainly does not look like an acceptable version. There are several possible reasons for this, all of which are related to the adjustments and operation of the oscilloscope. This should not be taken to mean that every trace, which is not a good version of whatever may be the kind of voltage being examined, is a sign of improper operation of the oscilloscope. It happens to be so in this example, the conditions of which have been deliberately chosen so as to illustrate several points. Once the proper method of setting up a trace on the screen has been learned, the proper and improper conditions of operation will be understood.

Returning to the reasons for the appearance of the trace on the screen (Fig. 10-48A), one of these is insufficient horizontal deflection; the horizontal-amplifier gain control is not advanced sufficiently relative to the amount of vertical deflection. A second reason is that the absence of absolute linearity in the sweep voltage tends to distort a single cycle. The third reason also is related to the sweep voltage. The distortion is partly due to the portion of the signal cycle occupied by the retrace or flyback time of the sweep voltage. Since this never can be sufficiently rapid so as to occupy zero time, it must cut off a portion of the cycle in the display.

With a single cycle on the screen, we can perform several experiments. The first of these would be described as "increasing the sweep amplitude." This is done by advancing the horizontal gain control somewhat, and this produces the trace shown in Fig. 10-48B. Actually, we have increased the level of the time-base voltage *at the deflection plates*, but *not* by increasing the sweep oscillator output; that remains fixed, for the output from the time-base generator *is constant* in level at each of the frequency settings. The output may not be the same over the entire range of the oscillator, but it is substantially constant over a given part of the range, tending to fall at the high frequencies. However, at any one frequency, any increase or decrease in time-base voltage at the deflection plates is a function of the horizontal gain control, since it determines voltage level of the signal admitted into the horizontal amplifier.

The amplitude is increased, as we stated, and the pattern (B) in Fig. 10-48 is a much better-looking wave than pattern (A). Bear in mind that no change has been made in the audio oscillator output, nor has any change been made in the vertical amplifier. It is, therefore, evident that the adjustment of the trace dimensions can have a very material effect on the general appearance of the trace, and can very greatly influence visual interpretation.

The improved pattern is still far from being a really good wave, since some of the conditions mentioned previously have not yet been changed. Increasing the amplitude of the sweep voltage by advancing the horizontal gain control does not improve the linearity of the sweep, because, as we stated earlier, the control is external of the sweep generator. It is located in the horizontal amplifier, and we assume that this system contributes nothing to, or detracts nothing from, the characteristics of the time-base voltage. This may not always be so, but it is regarded as being so at this moment.

The fact that the return trace, visible as a very thin and very light line across the screen, is not located at the exact center of the trace is a matter of the phase relationship between the synchronizing voltage and the time-base oscillating system. (This is discussed in Chapter 11.) Perfect centering so as to divide the trace into two halves never is attained, but a fair approximation is accomplished, although in the final analysis, it is not necessary, for reasons which soon will become evident. A better approach to centering of the return trace can be made by slight readjustment of the synchronizing signal amplitude control and the fine frequency control, but it is not too important.

A still further improvement in the appearance of the trace can be made by advancing the sweep amplitude (horizontal gain control) a little more, until the trace shown in Fig. 10-48C appears. If you compare (A) and (C), the effect of proper trace dimensions is very clearly seen. This does not mean that every trace should be spread horizontally as much as possible, for by so doing one can alter the *apparent* characteristics of a wave so that it will appear like something which it is not. Just what we mean will soon be shown.

How can we minimize the effects of a nonlinear sweep voltage? Since the nonlinearity is more pronounced as the maximum charging voltage point is reached in the sweep oscillator, one means of minimizing the effects of a nonlinear sweep is by placing several cycles of the signal voltage on the screen. The cycle or cycles which are spread by the linear part of the sweep will be properly shown, and the portions which are distorted by the nonlinear part of the sweep may be ignored. Such a condition is approximated when 3 cycles of the signal are shown on the screen. It would be even more true if more cycles were displayed, but as we said previously, crowding of the signal cycles in the display impairs the ease of examination.

So, we reduce the frequency setting of the sweep oscillator by means of the rough frequency control to about one-third of what it was before. Since the signal frequency is 750 cps, the time-base oscillator in the oscilloscope is adjusted to approximately 250 cps. The setting of the signal source remains untouched. In order to make the pattern stationary, the fine frequency control will require adjustment, as will perhaps the synchronizing control. The result is shown in Fig. 10-49A. If you compare this display with (A) in Fig. 10-48, the improvement effected by showing 3 cycles instead of 1 cycle is immediately evident. It is still the same signal, except now it looks very much better, and this is a truer representation because the conditions of operation of the oscilloscope are much more correct.

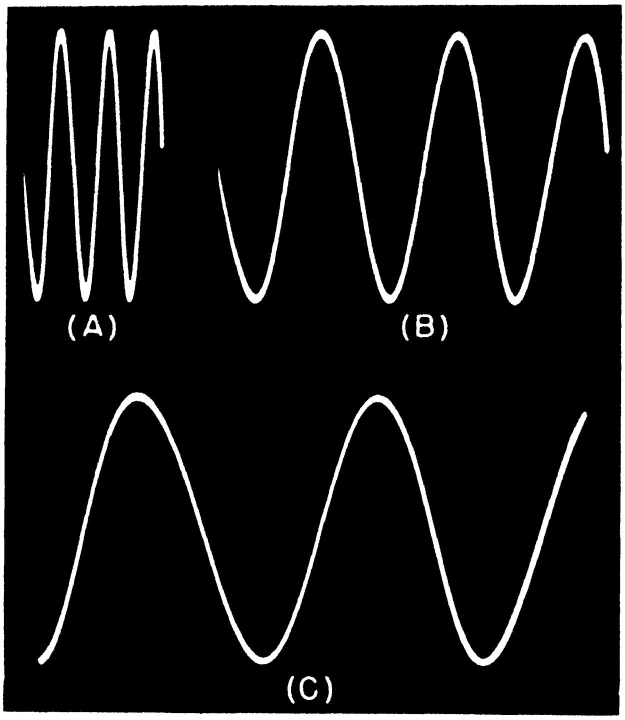


Fig. 10-49.—Illustrating the improvement in a 3-cycle pattern as a result of increasing the sweep amplitude.

In truth, these 3 cycles are crowded too much for easy examination. The horizontal dimension originally satisfactory for 1 cycle is too little for 3 cycles. So we increase the sweep amplitude by advancing the horizontal gain control a little more, and the result is shown in Fig. 10-49B. To say the least, these representations of the signal from the audio oscillator more closely approach the normal impression of a sine wave; the trace is of such proportions that the characteristics of the wave can be studied, especially the middle cycle.

If it is desired to spread the wave still more, the sweep amplitude (the horizontal gain control) is advanced still further and the pattern shown in Fig. 10-49C develops. It may seem at first glance that the frequency of the sweep was changed, because fewer cycles appear on the screen. What has actually happened is that the increase in sweep amplitude has increased the beam deflection in the horizontal direction sufficiently to cause it to move beyond the limits of the screen and those portions of the wave which are being traced by the beam at the extreme limits of its travel are no longer visible because they are off the screen.

Horizontal amplitudes of this dimension are not necessarily examples of incorrect operation. In fact, the reverse may be said under certain conditions. It is an example of good operation of the oscilloscope, when, because of the frequency limitations of the time-base oscillator, it is impossible to make the sweep frequency high enough relative to the signal frequency, to place 3 cycles on the screen. Under such conditions, the sweep amplitude is increased sufficiently so that only a few cycles appear on the screen, thus overcoming the handicap introduced by the limited sweep-frequency range, which, if treated otherwise, would crowd the screen with many cycles of the signal.

If you compare (A), (B), and (C) in Fig. 10-49, you will note that the peaked tops of the signal cycles in oscillogram (A) were due entirely to the conditions of display, for actually the tops of the wave are rounded off. It is in this connection that a word of caution is necessary; excessive expansion in the horizontal direction can round off the peaks of the wave being shown so that it becomes something which it is not. It is an action which must be remembered as a fundamental condition related to correct operation of the oscilloscope, and as will be shown soon, applies in the vertical direction too. Certain optimum dimensions exist, but they cannot be stipulated as definitely as we state the minimum number of cycles of a wave. Experience is the best teacher.

We might further remark that the display of 3 cycles eliminates all complications due to the loss of a part of a cycle by the sweep voltage retrace. Since this takes place on the last cycle and also influences the point where the display of the first cycle begins, the second cycle is shown completely without being affected in any way by the return trace time.

Vertical Amplitude Variations

The images shown thus far were made with the initial vertical-amplifier gain-control setting. Understandably, adjustment of this control can have its effects. Let us imagine a reduction in the gain made by reducing the setting of this control. What happens? The obvious result is the reduction in the height of the trace on the screen as shown in Fig. 10-50A, but in addition, such reduction in vertical-deflection voltage may be, and usually is, accompanied by the need for a slight readjustment of the synchronizing voltage control. This stems from the fact that reducing the signal input to the vertical amplifier reduces the amount of synchronizing voltage applied to the time-base oscillator. If the reduction is sufficient, the sweep oscillator will go "out of sync" and the trace will be set into motion. As the vertical-amplifier gain control is retarded, the synchronizing control must be advanced in order to apply the same amount of synchronizing voltage to the sweep oscillator. We assume that this was done and the pattern of Fig. 10-50A is the result.

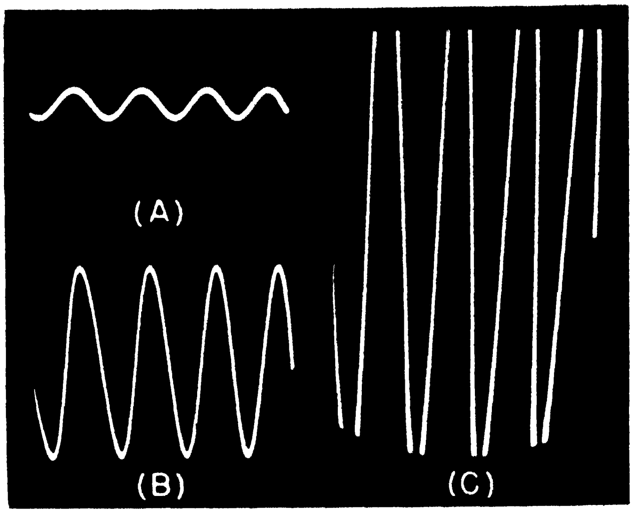


Fig. 10-50.—Showing the effect of varying the vertical amplitude of a particular signal.

It is easy to see that the trace is out of proportion, that is, it is too wide for its height. The wave still looks like a sine wave, but such reduced height can cloak some irregularities in the wave; at any rate, a height equal to the width is preferable. So the vertical-amplifier gain control is advanced and the trace shown in Fig. 10-50B appears on the screen. Depending on the setting of the synchronizing control, the wave simply increases in height without any major change in the number of cycles shown, although, if the synchronizing control is advanced too far, increasing the vertical-amplifier gain control may alter the number of cycles on the screen. This may require a slight reduction of the synchronizing voltage, which we assume has been done, in order to retair the same general appearance of the trace.

Advancing the vertical-amplifier gain control still more may result in a trace like that shown in Fig. 10-50C. The peaks of the wave are off the screen because the vertical deflection is excessive. A display of this type is not usually used, although it has its purpose if the primary concern is in the portions of the wave between the peaks. When the interest lies in the entire wave, such an adjustment of the vertical-amplifier gain control is incorrect.

If the interest lies in either the positive or the negative peaks of the wave and it is desired to examine them under conditions of high amplification, the beam-positioning control may be used to shift the entire image downward or upward so as to return the missing peaks onto the screen. This is shown in Fig. 10-51 A and B. Oscillogram (A) is the display when the beam-positioning control has shifted the beam sufficiently so that the positive peaks are visible, but almost the entire lower half of the trace is off the screen, and oscillogram (B) shows just the reverse condition. In this illustration, the positioning control has shifted the beam sufficiently so that the negative peaks appear on the screen, but almost the entire positive alternation of the wave is off the screen. Whether or not these variations in the settings of the vertical-amplifier gain control require readjustment of the synchronizing control depends upon the exact conditions in each instrument. If they are required, we assume that the slight changes necessary to stop the image are made.

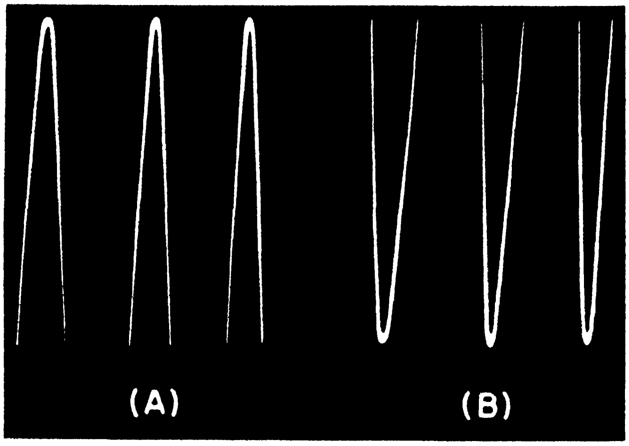


Fig. 10-51.—Positive peaks and negative peaks of a signal whose peaks would normally appear off the face of the screen.

Trace Blurred at Edges

It might be well to make several more comments concerning the vertical dimension of the image. To begin with, the usual method of operation is to set up a trace which is as large as the signal conditions and the screen dimensions allow, bearing in mind that the usable portion of the screen surface does not extend out to its boundary. If the peaks of the wave are permitted to extend to the extremes of the screen, the curvature of that surface will distort the trace at that point and make it difficult to examine. The distortion may be a change in the shape, but even if it is not so, the trace will be blurred. This is shown in Fig. 10-52A.

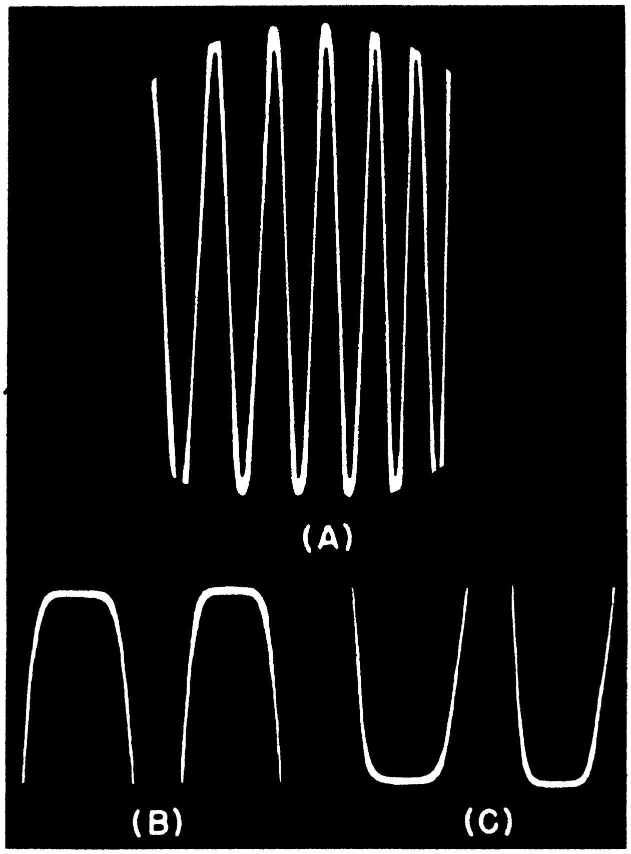


Fig. 10-52.—(A) Blurring at the edges of a pattern which extends to the extremes of the screen. (B) and (C) are the positive and negative peaks of a signal that results from overloading the vertical amplifier.

Overloading the Vertical Amplifier

Although the design of the vertical amplifier in the oscilloscope is such that it will accept a wide range of signal voltages, it is still possible to overload the amplifier by opening widely the vertical-amplifier gain control. Unlike the normal behavior of an overloaded r-f, i-f, or a-f amplifier in a receiver, which may indicate the condition by some departure from normal sound, overloading the vertical-amplifier tubes in an oscilloscope will not be audible. In fact, it may not even be visible in the display which appears on the screen.

Since the action is due to the high level of the input signal, the deflection of the beam will invariably be of such dimensions as to drive the beam beyond the limits of the screen or near the peaks, so that neither the positive nor the negative peak of the signal will be visible. Since the distortion takes place on the peaks, it can be detected only by adjustment of the beam-positioning controls so as to return the peaks to the screen, first the positive peak and then the negative peak.

The sign of such overloading is a flattening of the trace at the peaks of the wave as shown in Fig. 10-52B and C, the former being flattening at the positive peaks and the latter at the negative peaks. As a rule the vertical amplifier in an oscilloscope is designed so that the trace may be increased in size somewhat beyond the limits of the screen without overloading of the amplifiers. One might say that amplification beyond the limits of the screen is useless since what is beyond the screen limits cannot be seen anyway, but that is not entirely true, because the occasion may arise when it may be desired to expand the image vertically for a better examination of the waveform and to use the vertical positioning control to make the peaks visible. It is for such purposes that the amplifying capabilities of the vertical amplifier extend beyond the needs set by the screen diameter in the vertical direction.

As we mentioned earlier, it is difficult to designate the correct size of an image. It depends upon numerous conditions. Sometimes the gain of the vertical amplifier in an oscilloscope may not be sufficient to raise the amplitude of a very weak signal beyond a vertical dimension of 0.5 or 0.75 inch. Unless additional amplification is available between the signal source and the input to the vertical amplifier, one will have to be satisfied with an image of such vertical proportions. There are times when even with added amplification this amplitude is not available, in which case the oscilloscope is useless. When the height is small, the width must be reduced in proportion. In addition, it then becomes advisable to reduce the dimensions of the spot or the beam density so that a thin line trace is attained. It is easier to study the characteristics of such a small trace when portrayed by a beam of reduced density, than when the beam is at full density.

Insufficient and Excessive Synchronization

Although we have spoken about synchronization on several occasions, still more remains to be said. We still have in mind internal synchronization by the signal secured from the vertical amplifier, just as was indicated in Fig. 10-45, and we are still working with the sawtooth time-base voltage.

There are two conditions of incorrect synchronization, regardless of the origin of the synchronizing signal. One of these is insufficient synchronizing voltage, and here the pattern will be in motion; the other is excessive synchronizing voltage, in which case the pattern will be stationary, but it will be distorted, and in addition the frequency of the time-base oscillator may be materially changed. The origin of this distortion is very interesting, but, since it is treated in detail in the chapter which follows, we shall be brief at this time. The same applies to the control of frequency.

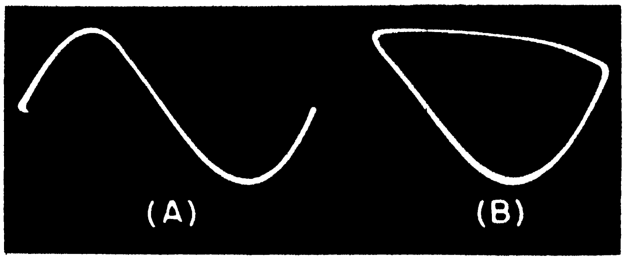


Fig. 10-53.—(A) Properly synchronized sine wave. (B) Same signal oversynchronized.

If we were to decide between two evils, that is, under- or over-synchronization, the latter is preferable since the image is stationary, but its adverse effects can be so great as to warrant the statement that it is no better than the first. For example, Fig. 10-53A illustrates a sine wave shown as a single cycle when the synchronization is correct. With excessive synchronization, the same pattern is changed to that illustrated in Fig. 10-53B. Although we do not recommend single-cycle display, the trace (A) is readable, and note that the setting of the synchronizing control is such that both halves of the cycle are almost fully displayed. The pattern shown in (B) is an extreme case of excessive synchronization, but it need not be due entirely to the adjustment of the synchronizing control; it can be the result of that adjustment plus the level of the signal in the vertical amplifier, for, after all, the signal supplied by the synchronizing circuit to the sweep oscillator is determined initially by the signal fed into the synchronizing system from the vertical amplifier.

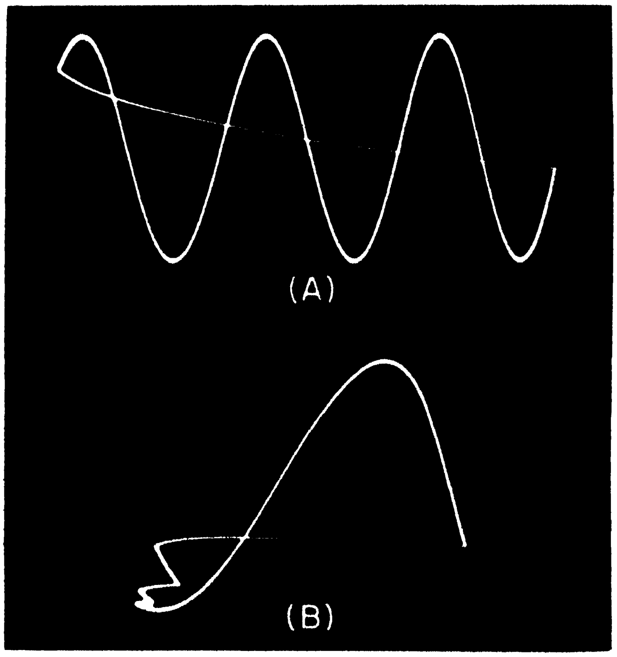


Fig. 10-54.—(A) Three cycles of a sine wave properly synchronized. (B) Same three cycles oversynchronized.

In order that we be consistent in the presentation, Fig. 10-54 illustrates (A) a proper signal of several cycles with correct synchronization, and (B), the effect of excessive synchronizing voltage on the same signal. Various stages between (A) and (B), representative of different degrees of synchronization, are possible, all of which would be illustrations of a gradual deterioration of the trace from (A) to (B). In fact, if the level of synchronizing voltage is high enough, a change in the sweep frequency sufficient to reduce the number of cycles from three to two is easily possible. This is explained and illustrated in Chapter 11.

Sweep Frequency Higher Than Signal Frequency

Let us now examine a different condition which is possible during operation. It is not the usual approach to waveform analysis, but it may be a temporary state while making the proper sweep-frequency adjustments. As a matter of fact, it is a clue to the condition where the sweep frequency is higher than the signal frequency. We are referring to a trace which seems to be lying on its side. One example of this appears in Fig. 10-55A, and another, which is much more frequently encountered, is shown as oscillogram (B) in the same illustration. Both are representative of a sweep frequency which is twice as high as the signal frequency. The difference in appearance of the two traces stems from the phase conditions between the synchronizing voltage and the sweep oscillator voltage, the former being the equivalent of the input signal. Two possible conditions of synchronization during the development of the trace shown as (B) in Fig. 10-55 are shown as (A) and (B) in Fig. 10-56. Both of these were in motion when the pictures were snapped. The former shows less synchronization than the latter. Virtually the same effects could be achieved by variation of the fine frequency control of the sweep oscillator when degree of the synchronization is maintained just below the point when the trace ceases its motion. If you will recall, the graphic development of the trace resulting from a higher sweep than signal frequency was given in Chapter 9, as in Fig. 9-7.

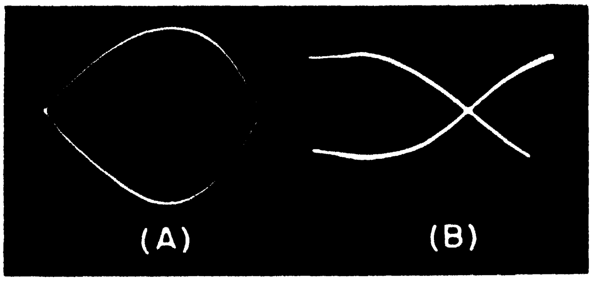


Fig. 10-55.—Sweep frequency twice signal frequency.

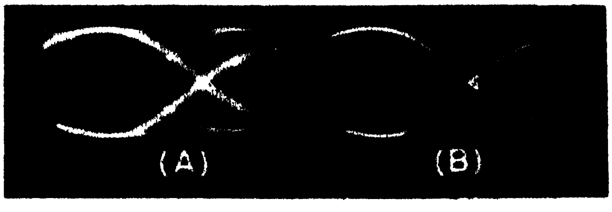


Fig. 10-56.—Stages in the synchronization of Fig. 10-55(B).

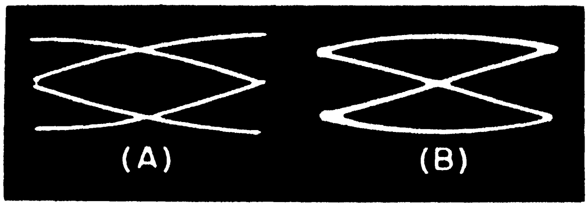


Fig. 10-57.—Sweep frequency four times signal frequency.

Another example of this frequency relationship, with a 4:1 ratio between sweep and signal frequencies, is given in Fig. 10-57A and B. The difference in appearance between them is simply a matter of the phase relationship between the synchronizing signal and the sweep voltage at the starting point. As a means of examining the waveform of the signal, these patterns are of no use, although under special conditions of control adjustments, these patterns may be used for calibration purposes. Much simpler and better means are available, however. All of the traces shown in Figs. 10-55 and 10-57 were made with the proper amount of synchronization. An idea of what can happen when the synchronization is excessive can be had from Fig. 10-58. Similar patterns are possible for higher frequency ratios between sweep and signal, but they serve little purpose because they become more and more complex with less and less meaning. Each line on the trace becomes a smaller fragment of the signal cycle. In Fig. 10-55, each line is a half-cycle, whereas in Fig. 10-57, each line corresponds to a quarter-cycle of the signal, since the horizontal movement of the beam is completed across the screen approximately once each quarter-cycle of the signal wave, excluding the retrace time.

The above illustrations deserve a few remarks. Synchronizing the sweep oscillator when it is set to a higher frequency than the signal seems to contradict the statement that the synchronizing voltage frequency must be higher than the sweep frequency in order to synchronize the latter properly. This is correct for reasons which are explained fully in Chapter 11, and the oscillograms of Figs. 10-55 through 10-58 are not contradictions. The ability to stop these patterns stemmed

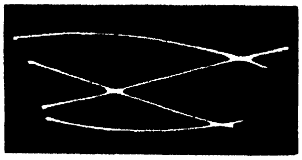


Fig. 10-58.—Fig. 10-57(B) with oversynchronization.

from the manner in which the equipment was set up for photographic purposes. Normally it is difficult to attain such long-period stationary patterns. Constant readjustment of the synchronizing control usually is necessary, or constant readjustment of the fine frequency sweep control is required. The action with pulse-type signal input is different.

Distortion of Complex Wave Input

The illustrations offered in connection with the display of a wave were accomplished with sine-wave input to the vertical-deflection plates. A complex signal can be treated the same way. This is shown in Fig. 10-59 for a properly synchronized input signal which is representative of a distorted voltage wave. By reference to Chapter 20, the wave shown was analyzed and found to contain the fundamental, third, and fifth harmonics; the fundamental and third harmonic being in phase, the fifth harmonic leading by 90°, and the amplitude of the third harmonic equal to 5 per cent of the fundamental, and the fifth-harmonic amplitude being about 12 per cent of the fundamental. Imperfect synchronization of the wave resulted in the oscillogram of Fig. 10-59B.

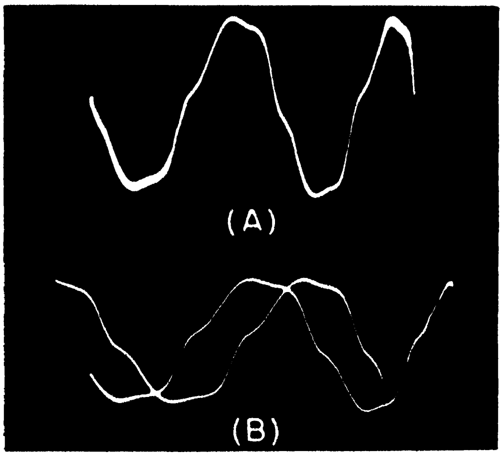


Fig. 10-59.—Complex wave (A) properly synchronized, (B) imperfectly synchronized.

Sweep Frequency for Complex Waves

The means of portraying complex waves which contain a number of frequencies may cause some doubt, especially since the sawtooth sweep voltage bears a single frequency rating. What is the frequency relationship between the sweep frequency and the signal frequency when the latter contains harmonics? Is the sweep frequency set relative to the fundamental of the complex wave or its harmonics? If so, which one? The answer is simple. Any wave which contains frequencies which are harmonically related to each other is displayed by selecting the sweep frequency relative to the fundamental frequency of the signal. By proper synchronization, the signal can be made stationary on the screen no matter how numerous its harmonic components. Such is the case in Fig. 10-59, where the sweep frequency is one-half of the fundamental frequency of the signal.

If the signal contains random frequencies, the sweep frequency should be selective relative to the lowest frequency in the signal, in which case those components which are harmonically related to the lowest frequency will be stopped, but there will be motion of the other frequencies through the pattern.

In the case of amplitude-modulated signals, the wave will be stopped if the modulation components are harmonically related and if the sweep frequency is selected relative to the lowest component of the modulation frequency. If, however, the modulation frequencies are representative of speech, the absence of a fixed fundamental frequency prevents a stationary pattern because the modulating frequencies are continuously varying. For sine-wave modulation (single frequency) the pattern is stopped just as if it were a conventional sine wave of the same frequency, by setting the sweep frequency for the modulation component.

The individual cycles of a sinusoidally modulated wave may be displayed if the sweep frequency is high enough, or if the carrier frequency is sufficiently low, by selecting the sweep frequency relative to the frequency of the carrier. Details of transmitter testing are given in Chapter 18.

External Synchronization of Sawtooth Oscillator

In all of the discussion which accompanied the explanation of waveform display with the basic oscilloscope, the synchronizing signal for the sweep oscillator was obtained from within the vertical amplifier. Provision for obtaining the synchronizing signal from the source different from the vertical amplifier in the oscilloscope also exists on the device. This is not used when the vertical amplifier feeds the signal to the deflection system, since then automatic supply of the synchronizing signal to the sweep oscillator is accomplished. However, since the facility exists, its use might just as well be demonstrated. It calls for setting the synchro-

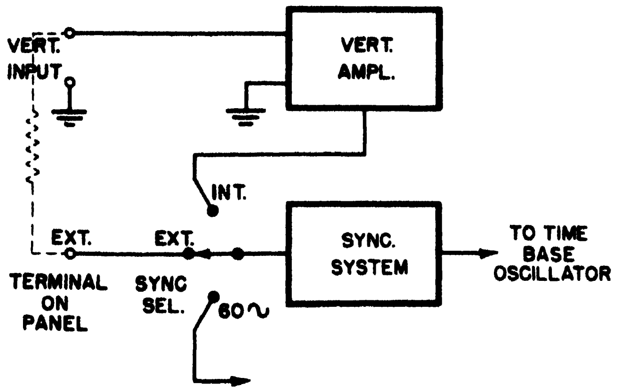


Fig. 10-60.—Sync selector switch set for external synchronization.

nizing system control switch to EXT (external) and connecting a lead from the source which supplies the signal to the vertical amplifier to the terminal marked EXT in the synchronizing system. This is shown in Fig. 10-60 by the dotted lines.

Making such a connection is simple, but several conditions must first be taken into account. The first is the level of the signal at the output of the signal source to which the external synchronizing circuit is connected. It may be sufficient for proper synchronization, but it is possible that it is too low (which seldom is the case), in which event recourse must be made to internal synchronization.

The second important point is that the input impedance of the synchronizing circuit usually is appreciably lower than that of the vertical amplifier; this may or may not influence the performance of the device which is supplying the signal to the vertical amplifier when the external synchronizing system also is connected to the signal source. If the former is the case, the external synchronizing circuit must be used, a high resistance of from 100,000 to as high as perhaps 250,000 ohms might be connected in series with the lead between the signal source and the EXT synchronizing terminal. Sometimes this series resistor is not required. This means that adequate signal level must be available at the output of the signal source. It then becomes a matter of determining which of these systems is permissible by virtue of the existing conditions, even including the use of the external synchronizing system as a whole.

Assuming that the conditions allow the use of external synchronization, how does this affect the performance of the oscilloscope? If adequate synchronizing signal is available, or rather if control of the synchronizing signal is correct, then the behavior of the oscilloscope remains as described. The only exception is that changes in vertical amplitude will *not* cause any effects in the sweep oscillator, for now the synchronizing signal is *not* varied by changes in the setting of the vertical-amplifier gain control.

Relative to the actual connection to the EXT synchronizing terminal, it joins the high side of whatever is the source of the synchronizing signal, or that point where an adequate signal will be available. Sometimes the oscilloscope may utilize a separate ground terminal in addition to the "high" terminal for the input to the synchronizing system. Whether or not this ground connection is needed depends upon the existence of a prior ground between the signal source and the oscilloscope. If such a ground exists, another ground is unncessary since all the grounds are common in the oscilloscope. Still another point relates to the usual series blocking capacitor which is present in the high side of the input to the synchronizing circuit. This capacitor bears a fixed breakdown voltage rating, and while in the most instances it is adequate, the voltage between the high point of the source of the synchronizing signal and ground deserves some thought.

60-cps or Line Synchronization of Sawtooth Oscillator

Among the synchronizing signals which are made available in the general run of oscilloscopes for the control of the sawtooth time-base oscillator is a voltage from the 60-cps power supply line. This can be used for the observation of signals which are integral multiples or submultiples of 60 cps. It should be understood that this is not the sine-wave time-base voltage, rather only a signal which will synchronize the time-base oscillator within the unit to 60 cps or to a submultiple thereof. Restated, this means that the frequency control switch governing the time-base oscillator must be set to the same frequency or to an integral submultiple of 60 cps. The sync selector switch position is shown in Fig. 10-61.

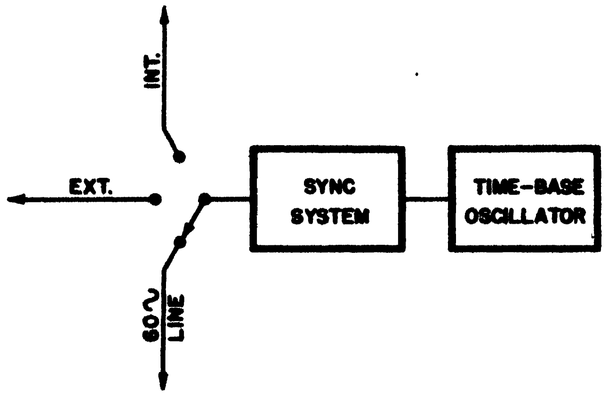


Fig. 10-61.—Sync selector switch set for 60-cps synchronization.

Distortion Due to Nonlinear Sweep

Although the matter of distortion of the trace due to nonlinearity of the sweep is really not a part of the fundamental applications of the basic oscilloscope to waveform display, it might well be mentioned, for it is entirely possible that some devices may show this characteristic at the high-frequency end of the sweep oscillator range. It may be due to nonlinearity in the sweep oscillator, or it may be due to the behavior of the horizontal amplifier when operated at this part of the sweep-frequency range. A very significant detail relative to the performance of the horizontal amplifier is the design of these systems so as to compensate for nonlinearity in the sweep, as described in Chapter 9, Fig. 9-102. Be that as it may, the presence of a sweep oscillator with such output will materially influence the shape of the trace on the screen. This is shown in Fig. 10-62. Oscillogram (A) is that of a substantially linear sweep and (B) is the resultant pattern of a sine wave developed with the sawtooth sweep shown. The reasons for several cycles of the sweep voltage is that it was separately photographed from the screen of another oscilloscope.

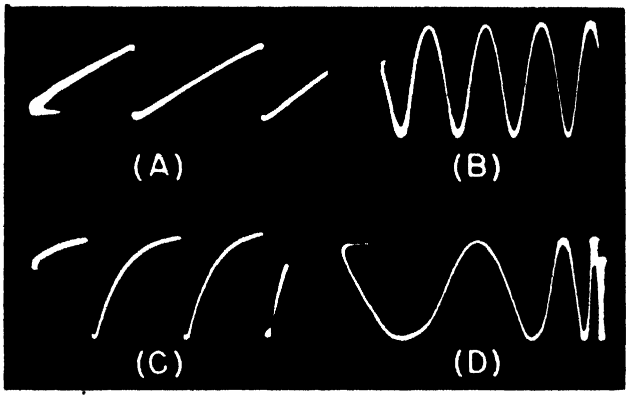


Fig. 10-62.—Distortion due to a nonlinear sweep. (B) is the sine wave produced by the linear sweep (A); (D) is the distorted sine wave produced by the nonlinear sweep (C).

The change in shape of the same sine wave when spread by a time-base voltage corresponding to oscillogram (C) (Fig. 10-62) is shown in oscillogram (D). The relatively small linear portion of the wave spreads one cycle of the wave fairly well, but after that the nonlinear portion crowds the remaining cycles. (This was explained graphically in Chapter 9, Figs. 9-18 and 9-22.) Obviously, distortion of this kind is not conducive to proper use of equipment.

Summary of Waveform Display

So much for the present about the use of oscilloscope controls when displaying waveforms. It is not the complete story, there being much more to come, but for the present it is sufficient to produce the necessary familiarity with the fundamental applications. A great many more of varied sorts are described in the pages which follow; the sole purpose of the present chapter is to indicate the functions of the controls. Many conditions may arise within an oscilloscope which are not bounded by the action of the controls. These will be brought to the fore after we have completed the sectional organizations of the basic oscilloscope and the explanation of the factors which govern the suitability of an instrument.

Sine-Wave Time-Base Presentation

Some applications of the oscilloscope require a sine-wave time base in place of the linear sawtooth. The examination of resonance curves or response curves of many varieties as in a-f, i-f, and r-f response and alignment, falls within this category. The sawtooth oscillator has been replaced by the sine-wave source, and the synchronizing system is omitted. All of this is accomplished by proper manipulation of the different selector switches as shown in Fig. 10-63. It may seem that the omission of the synchronizing system would interfere with the development of an apparently stationary pattern on the screen. That is not so when the signal under observation (applied to the vertical amplifier) varies at the proper time rate. This rate must correspond to either an integral submultiple or a multiple of the time-base frequency.

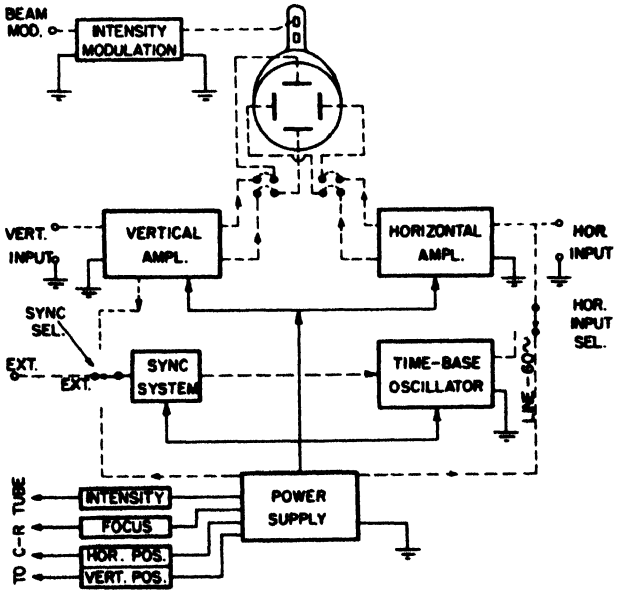


Fig. 10-63.—Block diagram of the basic oscilloscope arranged for 60-cps sine-wave time-base display.

Relative to these operations, some sweep generators supply the proper sweep signal, in which event the sine-wave sweep in the oscilloscope is not used. This is discussed in Chapter 16. The organization of the basic oscilloscope for such work is shown in Fig. 10-64.

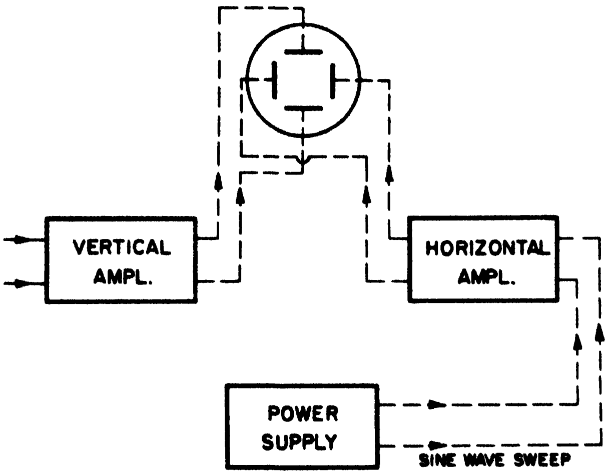


Fig. 10-64.—Schematic of the basic oscilloscope arranged for 60-cps sine-wave time-base display.

When the above is true, the pattern on the screen will appear stationary, but whether or not it will be interpretable for the desired information depends in great measure upon the specific situation. Usually, the design of systems which make use of the sine-wave time base is such that the amplitude variations of the signal fed to the vertical plates takes place at either the same rate as the frequency of the sine-wave sweep, or at twice that rate. Examples of such systems are fully described in Chapter 16. Just so that we may have some idea of the nature of traces developed in this manner, a typical response curve is shown in Fig. 10-65; this will appear on the screen when the vertical-deflection voltage varies in amplitude (in accordance with circuit performance) at a 60-cps rate, and the time base is the 60-cps sine-wave power-supply voltage. The curve depicts the behavior of a circuit over a range of frequencies. The signal fed into the circuit under test originates in a frequency-modulated oscillator, wherein the frequency changes at a 60-cps rate. The sine-wave time base in the oscilloscope advances the spot horizontally across the face of the screen in exact synchronism with the change in frequency in the f-m signal source. Since the time rates are fixed by the line frequency in both devices, the resultant pattern remains stationary on the screen.

It should be understood that the purpose of each of the two amplifiers in Fig. 10-64 is to raise the signal level to the value necessary for proper deflection and to afford control of the level for whatever sized image is desired on the screen.

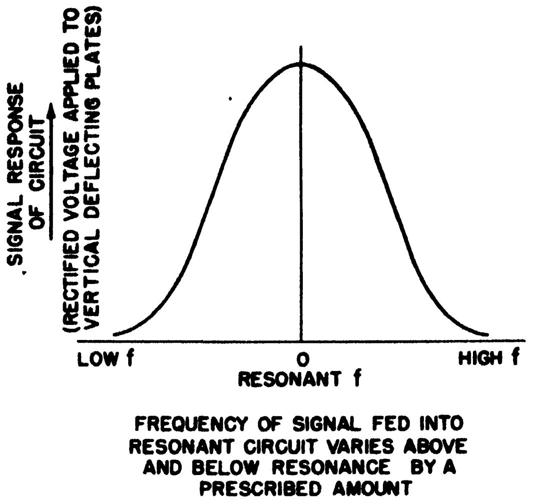


Fig. 10-65.—Typical single-peaked symmetrical response curve.

FACTORS CONTROLLING APPLICATION OF THE BASIC OSCILLOSCOPE

On several occasions earlier in this chapter, we stated that the performance specifications of a device determine the utility of the instrument. This is true for all kinds of oscilloscopes, not only the basic variety. These must be analyzed more completely from two viewpoints, the device proper and the systems to which the oscilloscope is applied. The two are interrelated; the performance specifications establish the capabilities of the oscilloscope, but even the best of these does not mean that it complies with the requirements.

Frequency as a Controlling Factor

Frequency response is the most dominant factor of utility. Virtually every section of the oscilloscope is beset by some sort of frequency limitation. Some of these are more pertinent than others, but all must be taken into account. Some of them offer greater latitude in application than others and these must be indicated. Let us analyze them individually and examine the manner in which they exert their control.

Impedance of Deflection-Plate System

A finite impedance exists between the two plates of each deflection system. It represents the capacitance between the two plates and the leads which feed the signal from the plate terminals to the plates. As such, each deflection system constitutes a frequency-resistive load on the circuit across which it is connected. If this circuit is the amplifier inside the oscilloscope, the capacitance of the deflection system and all of the connecting leads thereto are of little significance, because their effect is already taken into account in the published characteristics. But if the deflection systems are used with external amplifiers, the deflection plate capacitance and the capacitance presented by the circuits necessary to connect the amplifier and the deflection plates must be taken into account. Of these two sources of capacitance, the deflection plates are the least important, and may be neglected except at very high frequencies. But the capacitance presented by the connecting leads cannot be ignored; it can be sufficient to very materially influence the frequency response rating of the amplifier, especially at the high end, where it will reduce the response substantially, since it behaves like a shunting capacitor.

The temptation is to use a shielded lead to join the amplifier output and the deflection system. This is satisfactory if the frequency rating of the amplifier is predicated on the presence of the shunt capacitance presented by the connecting lead; if not, then the aforementioned effect will ensue. It is, of course, possible that the kind of work being done is such that the loss of a portion of the high-frequency response will have no ill effects. In this case, the entire matter requires no thought, but if the full bandwidth of the amplifier is required, then the combined capacitance of the deflection plates and the connecting leads deserves consideration.

The smaller the capacitance of these leads, the greater is the effect of the deflection plates, as small as that capacitance may be. As a rule, the latter is about 5 $\mu\mu f$. For the general range of frequencies, this is small enough and ineffectual, but when the signal frequencies exceed 1 Mc, even this small value of capacitance can prove troublesome.

When used with internal amplifiers, the deflection systems usually are shunted with fixed resistors of from 1.0 to 5.0 megohms, thereby placing the plates at substantially the same mean potential as the accelerating anode, and the deflection plate capacitance is in shunt with these resistors. When direct connection is made to the deflection plates, these shunt resistors are out of the circuit, and the input impedance rating stated in the specifications concerns only the deflection plates and the internal wiring.

Low-Frequency Response of Cathode-Ray Tube

The lower limit of frequency of the signals fed to the tube also has a bearing on the operation of the tube. The usual oscilloscope tubes use a P-1 screen, which has medium persistence. When excited by low-frequency phenomena, as, for example, between a fraction of a cycle and three or four cps, the persistence of the screen is too short to permit proper display. The screen is excited sufficiently, but the persistence characteristic of the phosphor results in a very rapid decrease in luminosity; thus the trace appears as an individual spot of light which moves, but does not show the path it traveled. No matter how frequently the spot retraces its path, it still appears as a single moving spot of light.

However, low-frequency signals may be displayed on a cathode-ray tube which has a long-persistence screen, such as the P-2 or P-7 variety. The afterglow displays the trace after the primary excitation has stopped; the action can then be studied. Thus, the low-frequency limitation may be overcome by replacement of the regular P-1 tube by one of like physical and electrical characteristics, but with a long-persistence screen.

Operation Features of Vertical Amplifiers

Several prime performance factors may be associated with the vertical amplifier in the oscilloscope. For the moment, let us view that amplifier as a "black box" connected between two points, *A* and *B* in Fig. 10-66. Point *A* is the origin of the signal voltage to be fed to the vertical-deflection system, and *B* represents the vertical-deflection plates.

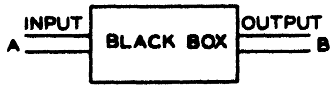


Fig. 10-66.—Vertical amplifier as a black box.

As to its function, that is self-evident; the very name assigned to the system describes what it does. Essentially, it amplifies a voltage. Depending upon the specific design, the character of the signals which it will accept may be of zero frequency (d-c) or high up into the megacycle region. Thus, it can be a d-c and a-c signal amplifier, or just the latter. Whichever it may be, it must bear a figure of merit which expresses the amount of amplification that it affords to a signal which is fed into its input at *A*, thereby expressing the level of the signal at *B*. This capability, or performance factor, is referred to simply as the *gain* of the amplifier.

Vertical-Amplifier Gain

While there is no standardization among amplifiers relative to gain, the very fact that similar devices are intended for similar functions and that they are sold at competitive prices establishes the general condition of similarity in groups. Such a situation is almost unavoidable, inasmuch as production costs are paralleled in the majority of manufacturing organizations. They may differ slightly, sufficiently, perhaps, to enable one organization to achieve performance factors slightly better than another; but generally it will be found that numerous points of similarity exist between the vertical amplifiers in the oscilloscopes which fall within the same utility category and within a certain price range.

Another reason which makes it difficult to assign specific values of gain in general terms is that it is very closely associated with another performance factor, namely, the bandwidth of the amplifier. This is the range of frequencies which the amplifier will accept and amplify uniformly. Although modern electronic design possesses great knowledge, it still has found no way of overcoming the design limitation wherein frequency bandwidth and gain vary inversely. This stems from the fact that both gain and frequency bandwidth are limited in each stage of the amplifier. While it is desirable to design a single stage which will afford very wide bandwidth, this is limited because of the distributed capacitances which are inextricably associated with each vacuum tube. Moreover, the circuit wiring also contributes to this effect, and while it can be minimized, it cannot be eliminated, so that even under the most favorable conditions, limitations exist. These are prominent at the high end of the frequency band, where the reactances of the shunt capacitances become small.

While it may not be evident, every attempt to raise the bandwidth reflects severe increases in cost of production; not only because the design is more critical, but also because the construction, too, is much more critical. Great care must be taken in such simple details as lead dress, because a change in the location of a wire relative to the chassis can cause an increase in the distributed capacitance. A change of this kind can adversely affect the bandwidth and the gain.

Still, it is necessary to quote average values of gain as found in the vertical amplifiers. While these figures may not be informative about all the instruments, they are useful as examples of what we mean by gain.

Take as the first example two different models of oscilloscopes intended for different applications, yet with like frequency bandwidths for the vertical amplifier, specifically, 100 kc. One of these affords a gain in voltage of 55, whereas the other affords one of 200. This ratio of about 4:1 in voltage gain over the same frequency band does not contradict the previous statements concerning the limiting effects of bandwidth. It simply means that a greater number of stages are used in the amplifier with the higher gain. Using the above gain figures as examples shows that it is impossible to arrive at an average gain or an average number of stages, unless we restrict the discussion to oscilloscopes that are within a certain price class. Then and only then is it possible to arrive at compromise values.

As another example, we might select an oscilloscope with a top vertical frequency of 2.0 Mc and a vertical gain of 250, or another with a top vertical frequency of 8.0 Mc and a vertical-amplifier gain of 125. These happen to be the products of a single manufacturer, but that is immaterial because it illustrates the point just as well as if we had chosen the products of different manufacturers.

We have indicated in the past that the vertical gain is associated with the deflection sensitivity of the cathode-ray tube used. Thus, if one tube has a larger screen than another, and/or there is a difference in deflection sensitivity, yet the two oscilloscopes using these tubes are intended for similar application (by which we mean that like levels of signal voltage may be anticipated at the input of the vertical amplifiers), then the gain of the vertical amplifier must be greater in one instance than in the other, in order to provide like vertically dimensioned traces on the screen.

Required Amplifier Gain

As to what should be the required gain for an application, that depends upon the values of voltage which may be expected from the sources under investigation and the amount of input voltage (at point *A* in Fig. 10-66) so as to provide at least 2-inch peak-to-peak signal on the screen. The usual statement of the gain capabilities of the vertical amplifier is in the form of an a-c input voltage in rms values per inch of deflection. Since an a-c voltage deflects the beam in both directions, the reference to the linear dimension of the trace means a *peak-to-peak* dimension.

This is illustrated in Fig. 10-67. The sine wave is shown as having an rms value of 0.8 volt. When such a value of voltage is identified as affording a 1-inch trace on the screen (without any horizontal displacement), the trace is a vertical line, 1 inch in height, as

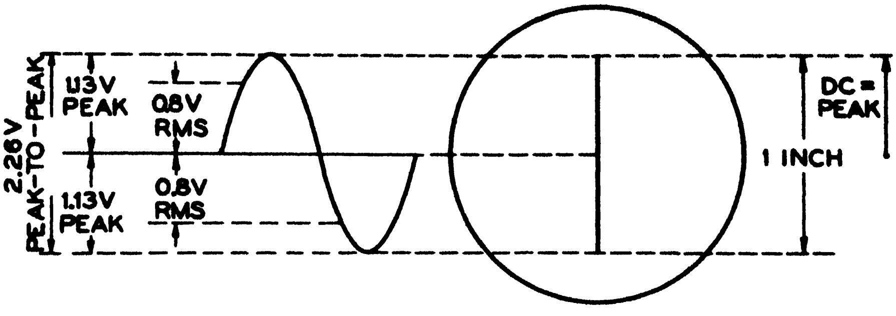


Fig. 10-67.—Relating rms input to peak-to-peak deflection voltage.

shown in the figure. Since the rms value is equal to 0.707 of the peak value on each alternation, any reference to an rms value automatically indicates a peak value equal to 1.414 rms, so that in the example chosen, the 1-inch trace equals a *peak* voltage of 0.8 × 1.414, or 1.13 volts on each alternation. Since the cathode-ray tube displays both alternations, the peak-to-peak value for a sine wave will be equal to twice the peak value of each alternation, which in the example chosen is 2 × 1.13 or 2.26 volts.

It has become standard practice to designate sensitivity of the complete vertical-deflection system in terms of the 1-inch over-all deflection, but, as previously mentioned, a deflection of such dimensions is too small for satisfactory observation; at least 2 inches are needed. This means that the sensitivity rating of the vertical system should be multiplied by two in order to establish whether or not the gain of the vertical amplifier is ample for the work to be done. This is not a firm rule, but it is our suggestion in the attempt to assure most convenient use of the oscilloscope. For example, if the lowest voltage from the source under investigation is 0.8 volt rms, according to our suggestion, the amplifier rated at 0.8 volt rms per inch would not be too satisfactory; for a 2-inch trace, that amplifier would demand an input voltage of 1.6 volts rms. The amplifier system which would afford a 2-inch trace with an input voltage of 0.8 volt rms would be one rated at 0.4 volt rms per inch. From our point of view, the ideal amplifier is one which affords a sensitivity numerically equal to *one-half* of a reasonable value of voltage available from the source under investigation.

Another reason which underlies our suggestion is that if the gain in the amplifier is greater than the minimum desired for operation, it is possible to reduce the gain, and in that way decrease the effect of undesired hum signals which may be present in the system being checked. Discounting for the moment the *possible* effects of attenuator adjustment, it always is advantageous to use less than the maximum available gain in an amplifier. A much cleaner trace usually results.

There is, of course, another side to the matter of high amplifier sensitivity. The more sensitive an amplifier, or the more gain that an amplifier provides, the more critical are the conditions of use. Such an amplifier will be more prone to pick up extraneous signals; its *input* will be more sensitive to improper grounding or shielding, to noise pickup, and to external fields. Under proper conditions of use, such pickup seldom is a problem, but under improper or careless conditions of use, the trace on the screen can be very greatly affected. This is why one oscilloscope may not show traces of undesired signals, whereas another with much greater sensitivity and used in the same place under identical conditions will indicate the presence of these fields.

In the case of amplifiers which have a frequency range extending down to zero, the conditions described above still apply. The fact that the amplifier response includes d.c. as well as a.c does not necessarily increase the sensitivity of the device because the fixed relationship between d-c and a-c voltage relative to the dimensions of the deflection remains unchanged. Refer to Fig. 10-67 once again. The peak voltage in *one* direction corresponds to a unidirectional d-c voltage of like amount. The peak-to-peak a-c voltage corresponds to an equal value of d-c voltage with the spot displacement calculated from one extreme of the a-c trace to the other, regardless of the frequency rating of the amplifier which passes the signal.

Summarizing the matter of vertical-amplifier gain, it really is not the complicated problem which may be the impression gained from this discussion. Admittedly, one must know the intended application of the device, especially the input signal voltages which may be encountered. Vertical-amplifier sensitivities between 0.05 and 0.5 volt rms per inch will be found satisfactory in most servicing applications. If the gain is higher, so much the better. However, the fact of cost cannot be denied; in this connection, a word of caution is in order. In the long run, the best turns out to be the cheapest; always buy the best that you can afford.

Frequency Requirements of Vertical Amplifiers

The second important capability of the "black box" in Fig. 10-66 is the frequency over which the amplifier will operate. Although not generally identified as such, every amplifier is, in a way, a filter system, because it will neither pass all frequencies, nor will it amplify equally those frequencies which it will pass. The pass-band of the vertical amplifier in an oscilloscope has a great bearing on the utility of the device. This is true, regardless of voltage waveshapes which will be fed into the amplifier, but most significantly relative to nonsinusoidal waves, the investigation of which is today one of the paramount fields of utility of the oscilloscope.

Equipment cost is very closely tied to bandwidth; the higher the bandwidth, usually the greater will be the cost of the device. One of the very significant changes which has taken place in the design of cathode-ray oscilloscopes during the past fifteen years has been a gradual increase in the bandwidth of the amplifiers in the device, especially the vertical amplifier. The need for this improvement originates in the waveforms which have become standard in many kinds of equipment. We are speaking about square, sawtooth, triangular waves, etc., or combinations thereof. Many modern devices and systems make use of such waveforms for operation or testing.

Importance of Harmonics

The amplification of a sine wave involves only one frequency at a time, and the frequency limits of an amplifier are dictated simply by the range of individual frequencies which are to be raised in level. It is not so when working with nonsinusoidal waves. The relationship between the fundamental frequency of a square wave, for example, and the frequency range of an amplifier is of some importance; of very much greater import, however, is the frequency range of the harmonics of that fundamental and the pass-band of the amplifier. In other words, an amplifier capable of amplifying all frequencies between, for example, 10 cps and 1.0 Mc within 3 db, could be used for the amplification of sine waves substantially beyond this frequency band, but it would not be suitable for the amplification of square waves having fundamental frequencies above about 100,000 cps.

The importance of the frequency response in the amplification of nonsinusoidal wave lies in the ability of the amplifier to uniformly amplify the harmonics as well as the fundamental, in that way affording good *fidelity* of signal reproduction. Then, for oscilloscope amplifiers, what shall be considered the required frequency range? Unfortunately, the answer is not as simple as the question. We know that the ideal condition never can be satisfied, because the ideal square wave, triangular wave, or sawtooth wave requires the presence of an infinite number of harmonics. This is not attainable because it would require an amplifier with infinite bandwidth. So a compromise is forced on everyone and it then becomes either a matter of individual preference, or the fulfillment of accepted specific standards as to what shall be considered good quality of reproduction.

Different applications or measurements dictate different conditions. An order of quality acceptable for one application is not necessarily equally acceptable in another case. This is why it is possible to generalize, but not to set definite minimum requirements. There is no agreement among engineers as to the requirements, although there is general concurrence concerning a range of the *minimum* number of harmonics which should be present for a *reasonable* quality of reproduction of a nonsinusoidal wave. The absence of sufficient harmonics, nonuniform amplification of the harmonics, or a radical shift in relative phase of the frequency components of a wave during amplification can produce a very badly distorted reproduction of the input signal. This may be sufficient to destroy the utility of the device as a tool for study and comparison.

It would do no harm at this time to devote some attention to the composition of the varieties of nonsinusoidal waves previously mentioned. These are but a few of the great many which can be illustrated, but they will serve their purpose because they will convey the impression so necessary for the understanding of what will follow.

Composition of a Square Wave

In Fig. 10-68 is shown a conventional square wave, with like positive and negative alternations. The letter E designates the amplitude of the wave. The wave shown is ideal, which means that the rise from zero to the maximum amplitude takes place in zero time. Prac-

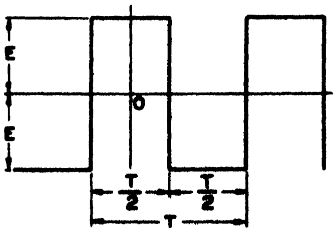


Fig. 10-68.—Conventional square wave.

tically speaking, this is impossible. In such a wave, the corners are perfectly square and the constant amplitude of the positive alternation and the negative alternation is shown by a perfectly straight line. Whatever the value of E, the voltage rises instantaneously to maximum, remains there for a while, and then falls in a straight line through zero to the maximum amplitude in the negative direction. This may be restated as saying that both the *rise time* and *decay time* are zero.

The period in which a cycle of amplitude change takes place is T, expressed in seconds; the reciprocal of this time, or $1/T$ corresponds to the *fundamental* frequency of the wave in cycles. If, for example, T is 0.0001 second, the fundamental frequency of this wave is 1/0.0001 or 10,000 cps.

The means of developing a square wave are numerous. It may be the simple expedient of controlling a d-c source, it may be the clipped output of a multivibrator, or it may be the process of successively amplifying and clipping a simple sine wave. While the fundamental frequency of such waves may be obvious because of the apparent association with time, the frequency requirements for proper reproduction are more obscure. The fact that a square wave, or any nonsinusoidal wave, may be resolved into a number of harmonically related sine waves, each of a certain amplitude and relative phase to form a square wave or some other nonsinusoidal wave, is a mathematical approach. This is the Fourier analysis whereby a wave is broken down into its components.

Let us select as our example a sine wave of 1,000 cps, shown in Fig. 10-69A. The choice of frequency is arbitrary; what will be said and shown applies to any frequency one may choose. We shall assume that the amplitude of this wave, which is the fundamental frequency, is unity. The starting point of the wave is at the time axis and it advances in the positive direction. We mention this because there is another starting point which is used in the usual mathematical equation for a square wave. This point is 90° later, at the point indicated by the projection from the time base.

We will now add a third harmonic $f3$ in phase with the fundamental $f1$ and with an amplitude equal to one-third of the fundamental. This is shown in Fig. 10-69B; the fundamental and the third harmonic are illustrated by the light lines, and the resultant by the heavy line. The ideal square wave is shown by the dotted lines. Several pertinent details warrant mention. A comparison of the sides of the resultant wave composed of $f1$ and $f3$, and just the fundamental $f1$, discloses that adding the third harmonic very materially steepens the sides of the wave. A second effect is a modification of the top of the wave on the positive and the negative alternations. The area included by each alternation has been increased and the period representing maximum amplitude has been lengthened.

Let us add the fifth harmonic $f5$ to the resultant of the fundamental and the third harmonic. This is done in Fig. 10-69C; the added harmonic has an amplitude equal to one-fifth of, and is in phase with, the fundamental. The result is shown by the heavy-line curve and a comparison with the dotted-line square wave illustrates the closer approach to the desired ideal. The sides of the wave have been made steeper and while the number of ripples on top have been increased, the area included by each alternation also has been increased and the maximum amplitude condition prevails for a longer time; the top is flatter than before.

The addition of the seventh harmonic $f7$, equal to one-seventh the amplitude of, and in phase with, the fundamental is shown in Fig. 10-69D. An improved wave with steeper sides and flatter top is the result. We need not picture any more harmonics because what we have shown is ample to develop the proper line of thought. It is clearly evident that just as the addition of three consecutive odd harmonics to the fundamental transformed the simple sine wave into an approximation of a square wave, the further addition of a sufficient number of odd harmonics would cause very steep sides and a flat top.

Note that as the number of harmonics added to the fundamental increases, the magnitude of the amplitude variations at the top of the wave decreases. With a sufficient number of odd harmonics present, the ripples would approach an infinite number and form a straight line. Two other very significant details can be seen from Fig. 10-69, namely, that as the wave contains more harmonics, the junction of the sides and the imaginary axis approaches a right angle; also that as the harmonics in the wave are increased in number, the curvature of the sides of the wave at top becomes smaller, ultimately forming a right angle when the number of harmonics contained in the wave is infinite. From the practical viewpoint, these ideal conditions are reached much sooner; that is, as far as the eye can see, and equipment can perform, a good square wave is reached with relatively few harmonics.

Two other details deserve attention. One of these is the phase relation between the component frequencies; the other is the fixed amplitude relationship. Although the square wave of Fig. 10-69D is only an approach to a square wave, it nevertheless enables us to reach some conclusions. The phase is a fixed quantity based upon the mathematical analysis of such a

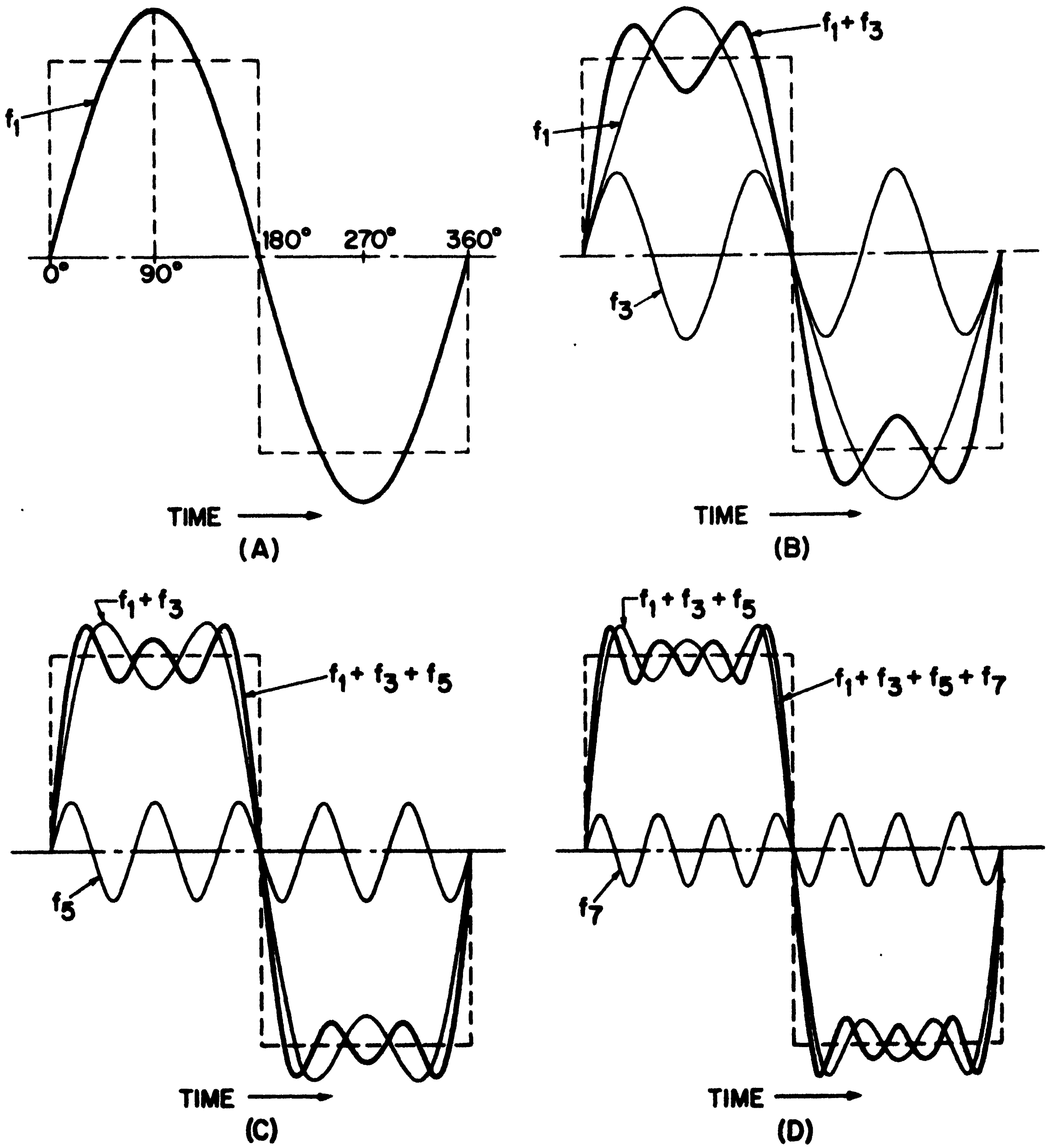


Fig. 10-69.—Composition of a square wave from its sinusoidal components.

wave. In the example chosen, we used in-phase sine-wave voltages. Exactly the same result will be obtained by the use of cosine waves, the point of origin of each component being 90° later. If you examine Fig. 10-69B, C, and D, and assume the starting point of each component as being located at 90°, as illustrated in Fig. 10-69A, it will be evident that the harmonics will alternately start in phase and 180° out of phase with the fundamental.

These references to phase lead to a very important conclusion. In our example, we showed the component phase conditions for the development of a square wave. In practice we are interested in the *reproduction* of a square wave which is fed into a circuit. Since a certain phase condition between the components is assumed to exist in the input voltage, it is *imperative* for proper reproduction that the *same phase condition between components exist in the output voltage*. This is the responsibility of the amplifier which is a part of the reproduction system; while it is amplifying the components of the wave, it must *retain* the phase relationship as well. A change in phase of a component or a group of components will very materially change the output waveshape. As you will see later, this is a matter of amplifier design, and also a matter of evaluating the constants of the amplifier which is used to amplify such waves.

Effect of Phase and Amplitude Changes

Fig. 10-70 shows oscillograms of the progressive stages of development of the square wave, which was illustrated graphically in Fig. 10-69. These were synthesized with the equipment used for the development of the waves shown in Chapter 20. Admittedly, the final wave does not have the shape which we would call even remotely acceptable as a square wave, but it nevertheless can be used for illustrative purposes and as a means of leading to certain conclusions. Since the first three odd harmonics have such a great effect, we can project actions associated with variations of these components into what may be expected under operating conditions when many more components are present in the wave.

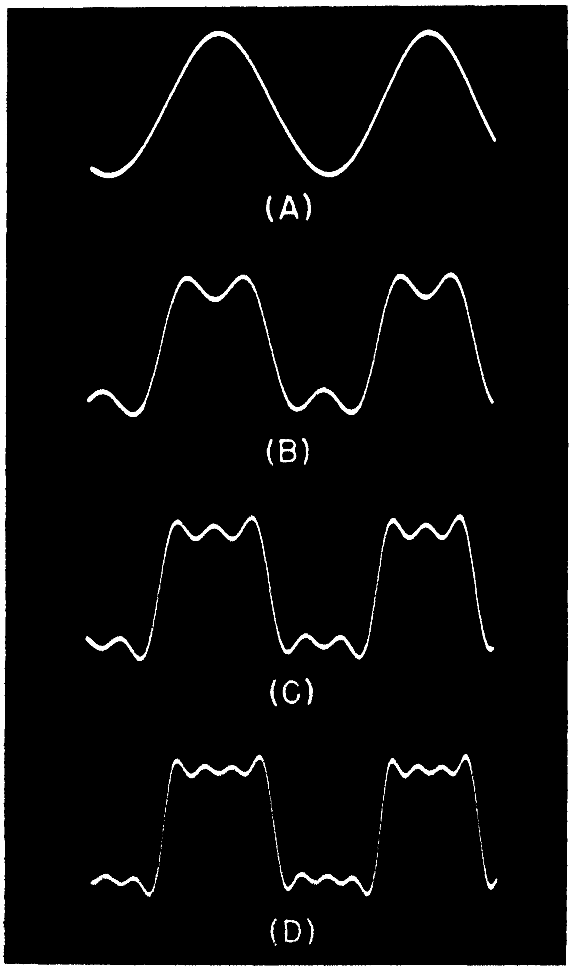


Fig. 10-70.—Oscillograms corresponding to the stages in the composition of a square wave of Fig. 10-69.

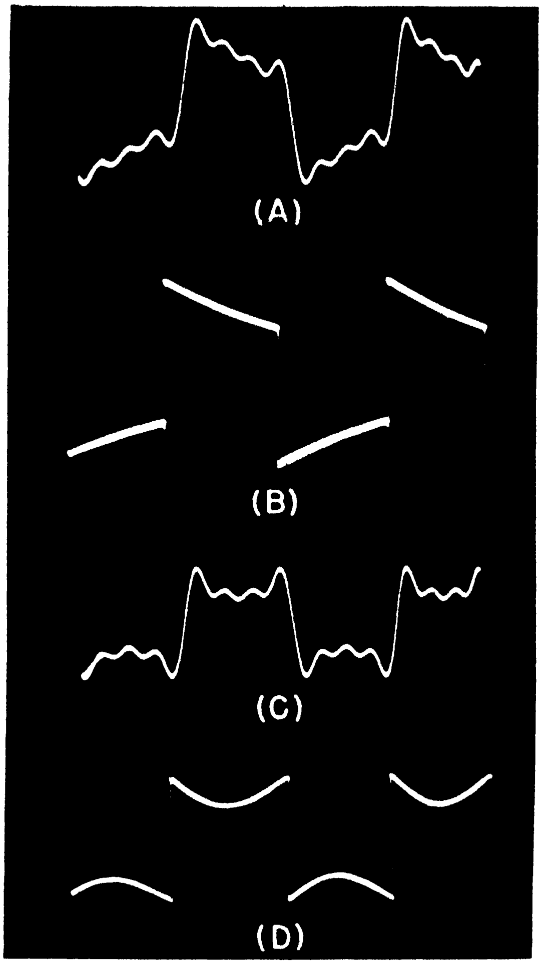


Fig. 10-71.—(A) shows the effect of changing the relative phase of the fundamental of Fig. 10-70(D), (B) shows the effect on the final square wave, (C) shows the effect of decreasing the amplitude of the fundamental of Fig. 10-70(D), and (D) shows the effect on the final square wave.

Fig. 10-71 shows the effects of a change in phase and a change in amplitude of the components. Curve (A) of Fig. 10-71 illustrates the effect of change (increase) in phase of the lowest component or the fundamental frequency. Although in the example cited, we have altered just one of the component frequencies, in actual practice, a similar change may take place to a varying degree upon a group of the lower-frequency components. The final result will be curve (B) of Fig. 10-71, although it may be much more aggravated than what we illustrate. A "tilt" has been created in the wave, and it often is expressed in an angular deviation from the ideal flat top of, for example, 8°, 10°, etc.

Such a situation is associated with the low-frequency behavior of an amplifier. Although most comments are made to direct thinking toward the higher frequencies, the lower frequencies are also important because they have tremendous effect on the range of fundamental frequencies which may be accommodated by an amplifier. Many amplifiers may afford the necessary high-frequency response to reproduce a square wave properly, provided that the fundamental frequency is not too low. If the fundamental frequency of the contemplated nonsinusoidal wave input signal is too low with respect to the pass-band of the amplifier, excessive phase shift or delay at the low frequencies will prevent the proper reproduction of the wave at the output of the amplifier, despite the fact that all of the required high frequencies are present in proper amplitude and phase.

Curve (C) of Fig. 10-71 shows the effect of a *change* in the *amplitude* of the fundamental frequency, that is, a reduction from the 100 per cent level which establishes the relative amplitude of the components. Of course, the converse of this would have the same effect, namely, excessive amplification of the higher-frequency components. Since it was easier to show the effect in terms of the lower frequency, it was used. The final result as it would appear if a group of the lower frequencies were attenuated, or a group of the higher-frequency components were accentuated, is shown in Fig. 10-71D.

Expressed mathematically, the equation for the square wave as developed in Fig. 10-69 is

$$e = \frac{4E}{\pi}\left[\sin \omega t + \frac{1}{3}\sin 3\omega t + \frac{1}{5}\sin 5\omega t + \dots + \frac{1}{2n-1}\sin (2n-1)\omega t\right]. \qquad (10\text{-}1)$$

However, the conventional mathematical analysis is based on the origin as shown in Fig. 10-68, which corresponds to the 90° point of Fig. 10-69A. Therefore, the *usual* form of the equation is

$$e = \frac{4E}{\pi}\left[\cos \omega t - \frac{1}{3}\cos 3\omega t + \frac{1}{5}\cos 5\omega t \dots \frac{(-1)^{n+1}}{2n-1}\cos (2n-1)\omega t\right]. \qquad (10\text{-}2)$$

In either expression only odd harmonics appear, which is the basic frequency condition not only for a square wave, but also for every wave which is symmetrical about the time axis. Symmetry means that the positive alternation and the negative alternation are alike to the extent that if the former is "flopped" over so that it is beneath the time baseline, it will be exactly the same as the negative alternation. The square wave and the triangular wave are examples of this.

Order Versus Number of Harmonics

It might be well to comment briefly about the difference between the presence of a certain number of odd harmonics and the significance of a certain order of the harmonic. Reference to three odd harmonics means the first three consecutive odd harmonics after the fundamental, or $f3$, $f5$, and $f7$; the numbers indicate the order of the harmonic. Thus the third odd harmonic is actually $f7$, equal to the seventh harmonic of the fundamental.

By examining equation (10-1) and the contents of Fig. 10-69, certain pertinent relationships become evident. While it is true that the first few odd harmonics contribute most to the squaring process, the higher harmonics cannot be ignored. The equation indicates that the amplitude of each component decreases inversely with the order of the harmonic. The frequency f_n of any order *odd* harmonic can be obtained from the equation

$$f_n = (2n + 1)f1 \qquad (10\text{-}3)$$

where $f1$ is the fundamental frequency and n is the order of the odd harmonic. Thus, the fiftieth odd harmonic of a 5,000-cps square wave is

$$f50 = (100 + 1)\ 5{,}000 = 505{,}000 \text{ cps.}$$

This corresponds to the 101st harmonic of the 5,000-cps fundamental. Its amplitude, according to the equation for the square wave, is 1/101 or 0.99 per cent of the amplitude of the fundamental. This may not seem much, especially in view of the relatively high amplitudes of the first ten or fifteen odd harmonics, yet it cannot be neglected completely.

In the case of a wave containing only even harmonics, the equation for the numerical value of even harmonic f_n, relative to the fundamental, is expressed by

$$f_n = 2nf1 \qquad (10\text{-}4)$$

where $f1$ is the fundamental and n is the order of the even harmonic. Thus if a wave contains 40 even harmonics, and the fundamental is 5,000 cps, the highest harmonic would be

$$f40 = 2 \times 40 \times 5{,}000 = 400{,}000 \text{ cps.}$$

The Importance of the Harmonics in Square-Wave Reproduction

The importance of a harmonic decreases with its order, but this does not imply that the addition of relatively few odd harmonics to a fundamental will produce an acceptable version of a square wave. In this respect, engineering practice has not yet decided upon what shall be the minimum harmonic content in a wave for

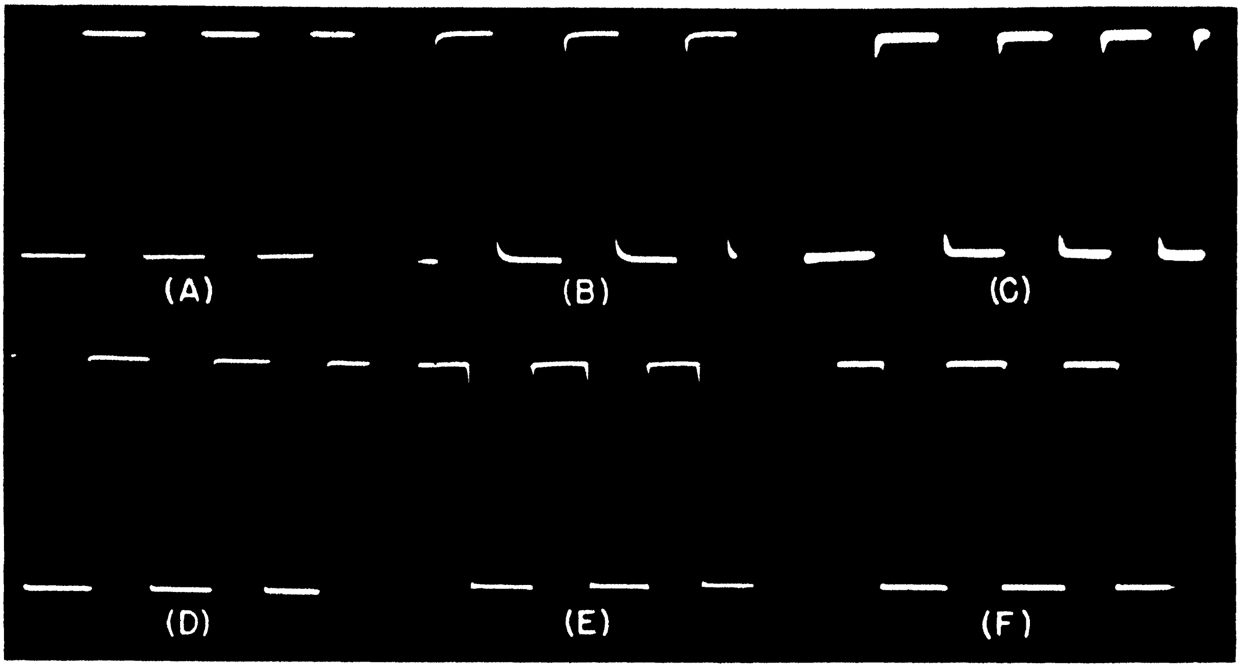


Fig. 10-72.—Square waves of different harmonic compositions; (A) is ideal, (B) contains 10 odd harmonics, (C) contains 25 odd harmonics, (D) contains 100 odd harmonics, (E) contains 165 odd harmonics, and (F) contains 500 odd harmonics.

the resultant to be an acceptable square wave. It is still a matter of preference and tolerable distortion.

However, some limit must be set, and while what we say may reflect general thinking, our comment is based on the broad aspects of the subject as related to maintenance activity. It is unfortunate, but it is nevertheless impossible to set clean-cut boundaries. For certain kinds of work, as few as ten odd harmonics of the fundamental will suffice, which means a frequency as low as twenty-one times the fundamental. Such a frequency range will result in rounded corners of the reproduced wave, but there will be enough of the flat measurement of amplitude and for general observation and analysis. This is commonplace in television servicing!

General television servicing, especially the *usual testing* of the pulse circuits in a television receiver, does *not* require more than ten *odd harmonics* in order to produce an acceptable representation of the waves found in the sync and sweep circuits. Admittedly there is a difference of opinion on this point; we are expressing our view. This does *not* mean that any system which affords such waveforms is suitable for the determination of *every* condition. A limitation in bandwidth is a limiting factor in application. Therefore if one desires more than the simplest in facility, then at least 40 *odd* harmonics of the fundamental should be within the amplifier pass-band.

The low-frequency response is also of importance in square-wave reproduction. A rule of thumb is that the lowest frequency square wave which will be satisfactorily passed by an amplifier is approximately ten times the low-frequency cutoff of the amplifier. Cutoff is defined as being 3 db below the flat portion of the response curve.

It is easy to see that if the desired condition is the presence of 10 odd harmonics, a 15,000-cps fundamental square wave requires a bandpass of 315,000 cps. The same bandpass would accommodate only about 5 odd harmonics of a 30,000-cps square wave, so that it becomes a matter of understanding the capabilities of the equipment. However, each individual or each application must determine what is considered an acceptable square wave after reproduction.

As a possible guide to thinking, concerning the composition of a reproduced square wave and the bandwidth of the amplifier, we show a series of oscillograms in Fig. 10-72. Curve (A) illustrates the shape of a 5,000-cps square wave when fed directly into the vertical-deflection plates of an oscilloscope. This is the standard wave for comparison; the period of the rise time is extremely small in proportion to the duration of the half-cycle. The remaining illustrations show the reproduction when (B) 10 odd harmonics are present, (C) 25 odd harmonics are present, (D) 100 odd harmonics are present, (E) 165 odd harmonics are pressent, and (F) 500 odd harmonics are present. In calculating the number of odd harmonics embraced by the amplifier pass-band, the upper-frequency limit was that frequency which was down not more than 3 db.

It should, of course, be realized that the pass-band has a top frequency equal to $(2n + 1)f1$, as previously shown in equation (10-3).

There is no doubt about an improvement in the waveshape as the result of adding higher-order harmonics, but there is doubt whether the added cost represented by the far more expensive design of the vertical amplifier is warranted; that is, determined by the utility of the information gathered from the oscilloscope screen. For ordinary amplitude measurements, the trace (B) in Fig. 10-72 is just as useful as the trace (F). These illustration show very clearly that the presence of the higher harmonics, with substantially uniform amplification, means faster rise times and sharper corners.

Pulse Duration and Rise Time

Although this text is not intended as an exposition of wave analysis, it is, nevertheless, necessary to mention another approach to the matter of the harmonic content of nonsinusoidal waves. The need for this is brought about by the apparent stress on the fundamental frequency of a square wave being the basis of determining the required pass-band of an amplifier which will pass such a wave. In instances when the square wave is symmetrical (i.e., equal positive and negative half-cycles), as in Fig. 10-73A where $d = T/2$, the fundamental frequency may be used as the basis of determining the required frequency response.

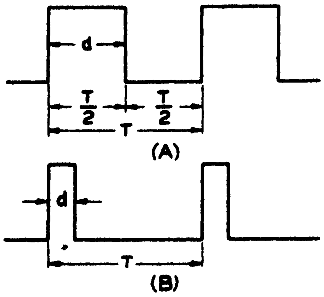


Fig. 10-73. — (A) is a typical square wave; (B) is an asymmetrical rectangular wave.

The asymmetrical *rectangular* wave of Fig. 10-73B introduces a new aspect. The rectangular pulses are of short duration d relative to the period T, and the rise time of the pulse is assumed to be small in comparison to the pulse duration. However, while $1/T$ is the fundamental frequency of the wave and the repetition frequency of the pulse, the determination of amplifier frequency response using the fundamental frequency as the basis would not be productive of a good version of the pulse.

If we assume that each pulse has a duration d of 5 microseconds, and T equals 1 millisecond, the fundamental (pulse repetition) frequency $f1$ is 1,000 cps. Based on the previous analysis, an over-all bandwidth which would accommodate 25 odd harmonics would be $(50 + 1) \times f1$, or 51,000 cps. Without going into the reasons why, such a bandpass would be totally inadequate for this waveform; this leads to the use of the *pulse duration* as the means of determining the determining the required pass-band. In this connection two approaches are used. The reason for there being two bases of calculating the bandwidth is another example of complete agreement on what represents an adequate reproduction of a square wave. One approach is the use of the equation

$$f_{\max} = \frac{1}{d} \tag{10-5}$$

where $f_{\max}$ is the required upper limit of amplifier response in cycles, and d is the pulse duration in seconds. For the waveform of Fig. 10-73B, this would mean a top frequency of $1/(5 \times 10^{-6})$ or 200,000 cps.

The other approach visualizes the required upper frequency in terms of the rise time of the pulse. In Fig. 10-73B, the rise time was considered small relative to the pulse duration. However, it was not taken into account on the supposition that, if the pass-band were adequate to permit the wave to rise to its full amplitude, the increased steepness of the sides which would result from increased bandwidth would not add to the utility of the pulse.

In view of the possible differences which may be introduced in the shape of this pulse by differences in rise time, it stands to reason that the rise-time consideration is the more important one if the shape of the pulse is the cardinal condition. It is only natural that the rise time be given consideration since zero rise time is an inmpossibility, and the shorter the pulse, the shorter *must* be the rise time if the pulse is to be an approximation of a rectangular shape. Fig. 10-74 shows an enlarged and exaggerated version of the pulse of Fig. 10-73B, so as to illustrate the meaning of the rise time. In conformity with standard practice, the rise time is the period between 10 and 90 per cent of the maximum amplitude. This period is assumed to be 1 microsecond.

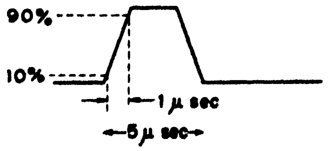


Fig. 10-74. — Enlarged and exaggerated version of a pulse of Fig. 10-73(B).

One school of thought establishes the required upper limit of the amplifier, in cycles, by using the equation

$$f_{\max} = \frac{1}{t_r} \qquad (10\text{-}6)$$

where t_r equals the rise time in seconds. For the pulse illustrated, the amplifier top frequency would be $1/(1 \times 10^{-6})$, or 1 Mc. Other authorities feel that an acceptable reproduction of the wave can be accomplished if the upper frequency is determined by making the numerator in the above equation equal to a value between 0.35 and 0.5, instead of 1.0. In the example cited, this means top frequencies of from 350,000 to 500,000 cps; in either case it is less than the 1-Mc bandwidth, which would afford a much steeper wavefront in the reproduced wave. Here is an example of great variance in bandwidth requirements in accordance with whatever is the school of thought on the subject and the desired accuracy of reproduction. Above all, it is an example of the use of factors other than the fundamental frequency as a means of determining the top frequency to be uniformly amplified.

Bandwidth in Television Oscilloscopes

In this connection, the requirements of oscilloscopes to be used for television servicing have been a subject of much discussion, because of the duration and rise time of the pulses that are examined during television receiver servicing. The usual rise time approximates 0.25 microsecond and the pulses may be as short as 5 microseconds. If the frequency needs of the shortest pulse duration are met, the required conditions for the longest pulse will be satisfied. If the frequency needs of the fastest rise time are met, the requirements for the longest rise time will be met.

Utilizing the approach of both pulse duration and pulse rise time results in amplifier response characteristics between a minimum of 1 Mc and a maximum of 4 Mc. From the viewpoint of the servicing industry, less than even the lower of these two bandwidths affords a satisfactory means of investigating the signals present in receivers. It has been found that a pass-band of 300 kc is adequate! Naturally, the lower-frequency response will introduce greater slope in the sides of the pulse and an increased rounding of the corners. Although such effects on the pattern *appear* to lower the utility of the low bandwidth amplifier, the fact remains that an oscilloscope with such an amplifier is *usable* for observation of even the short equalizing pulses as a part of maintenance operations. Of course, *greater* bandwidths afford a *better* trace and *improve* the shape of the pulses. So, again it is a matter of individual preference, and need versus cost. What is missing today is a definition of an acceptable reproduction of the square wave and its frequency composition.

Composition of a Sawtooth Wave

The use of the sawtooth wave is common in many applications which involve the cathode-ray tube. Consequently, its influence on the frequency response of the amplifiers which may be called on to amplify it is important. One would expect that only the horizontal amplifier should be a part of the discussion concerning sawtooth waves, because this time-base voltage is fed to the deflection system via the horizontal amplifier. However, the requirements are much broader because waves of this type may be fed to the vertical amplifier during the investigation of the output of one sweep system by means of another oscilloscope. It so happens that if the band-pass requirements of the amplifier are based on square-wave needs, they will adequately fulfill the needs of any other waveshape. However, as we mentioned on numerous occasions, increased cost is always a limiting influence. Therefore, every effort is made to keep costs down by reducing the pass-band. This is usually the case in the horizontal amplifier of the average oscilloscope. Since its primary function is to channel the time-base voltages, and since it is customary to show more than a single cycle of the signal under investigation, the frequency range of the time-base signal source is always less than the highest frequency which may be fed through the vertical amplifiers. If we add to this the reduced frequency needs of an amplifier which will reproduce sawtooth-shaped waves, we can understand why the bandwidth of the horizontal amplifier is usually much less than that of the vertical amplifier.

The mathematical equation for a sawtooth wave appears in two forms; one applied to the ideal wave shown in Fig. 10-75A, and one to the practical case shown in Fig. 10-75B. The fundamental difference lies in the time of the sudden fall from one extreme level to

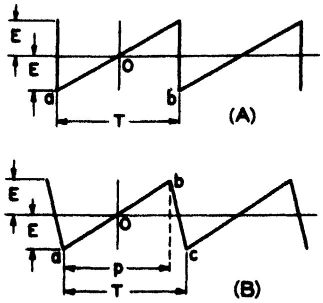


Fig. 10-75.— Sawtooth waves; (A) is ideal, (B) accounts for retrace time.

the other or *b-c* in curve (B). In practice this corresponds to the *retrace* time, or the time required for the beam to return from one end of its excursion to the beginning of the next trip across the screen.

The ideal sawtooth wave shown as (A) in Fig. 10-75 displays a zero time change between the peak and zero amplitudes, which is a condition unobtainable in practice. Under such theoretical circumstances, the useful portion of the wave is that shown between *a* and *b*, which corresponds to the period of one cycle of the wave. But, as we mentioned, the ideal cannot be achieved and we must accept a compromise wherein the retrace period involves a finite amount of time. This is the interval between *b* and *c* in wave (B). Although these voltage changes are nonuseful portions of the wave, they remain nevertheless parts of the cycle with *T* still the period of the wave.

Frequency Bandwidth for Sawtooth Waves

Let us examine briefly the composition of a sawtooth wave. This is shown in several steps from (A) through (G) in Fig. 10-76. These are predicated on the equation which assumes a flyback time equal to zero, the idealized condition, and is expressed as

$$e = \frac{2E}{\pi}\left[\sin \omega t - \frac{1}{2}\sin 2\omega t + \frac{1}{3}\sin 3\omega t \ldots + \frac{(-1)^n}{n}\sin n\omega t\right]. \tag{10-7}$$

The equation discloses a number of conditions. First, the wave contains both odd and even harmonics, in contrast to the wave symmetrical about the *Z*-axis which contains only odd components. Second, the amplitudes of these components vary inversely with the order of the harmonic. Third, the signs of the components alternate with the origin as the reference point. Nevertheless, like the square wave, the composition of a sawtooth involves a definite arrangement of components. If you examine the transformation of the sine wave as the harmonic components are added, the higher the number of harmonics contained in the wave, the steeper is the retrace portion and the less the period of this part of the wave. Also, as the number of components is increased, the useful portion of the wave approaches a straight line; the amplitude of the ripples becomes smaller, although the number of ripples increases.

The conditions associated with the number of components applies equally well to the nonideal sawtooth wave, which is expressed mathematically by

$$e = \frac{2E}{\pi^2 p^2(1-p)}\left[\frac{\sin \pi p}{1}\sin \omega t + \frac{\sin 2\pi p}{4}\sin 2\omega t + \frac{\sin 3\pi p}{9}\sin 3\omega t \ldots + \frac{\sin n\pi p}{n^2}\sin n\omega t\right] \tag{10-8}$$

where *p* is the portion of the cycle occupied by the linear rising portion of the wave. A significant difference is to be noted in how the magnitude of the components varies with the order. There is no simple expression for the manner in which the amplitudes of the harmonics decrease, since this depends on the value of *p*. For values of approximately 0.8 or 0.9 (10 to 20 per cent retrace), those components greater than about the fifth harmonic decrease much more rapidly than those for the ideal sawtooth. This is a very significant detail for the simple reason that it very greatly reduces the importance of the higher-order harmonics. In the idealized case, the eighth harmonic has an amplitude equal to 12.5 per cent of the fundamental, whereas in the practical case ($p = 0.9$), this amplitude is equal to about 3 per cent of the fundamental. The higher harmonics decrease even more rapidly.

The rapid decrease in importance of the higher order of harmonics has a major effect upon the frequency requirements of the amplifiers which channel such signals to the cathode-ray tube. It has been found that substantially uniform amplification over a range of frequencies with an upper limit equal to about 10 times the fundamental frequency of the sawtooth voltage is ample for the accomplishment of a reasonable and acceptable version of the sawtooth voltage fed into the amplifier. As in the case of the square-wave response, the greater the bandwidth of the amplifier, the more truthful the presentation, since it affords a greater number of harmonics in the reproduced wave. From the economical standpoint, a minimum of even seven harmonics is acceptable although at least ten should be present if the top frequency of a sawtooth voltage source is 30,000 cps, the amplifier which will channel this signal to the cathode-ray tube might well have uniform gain up to 300,000 cps. If the response is higher, so much the better; in the usual run of devices, it is definitely lower.

So far we have discussed two types of nonsinusoidal waves, the square wave and the sawtooth wave. By no means are these two types representative of the wide variety which may be experienced in practice. One of them, the square wave with its modification into a rectangular pulse is, however, representative of the most severe conditions which may be encountered. All others can be viewed as imposing simpler needs on the bandwidth of the amplifier. So, using the square or rec-

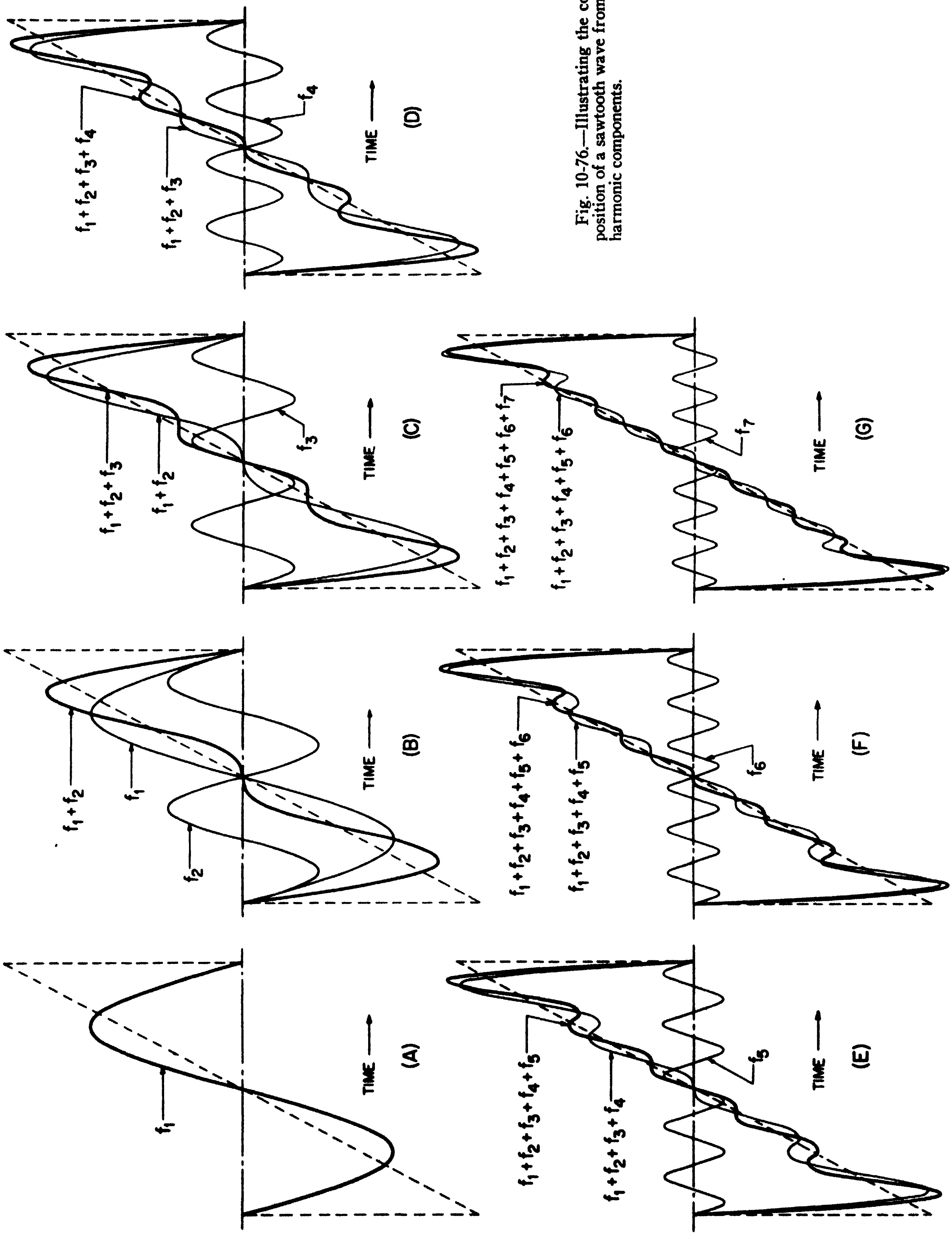


Fig. 10-76.—Illustrating the composition of a sawtooth wave from its harmonic components.

tangular pulse as the basis, we can summarize by saying that the steeper the wavefront, or the shorter the rise time, the greater is the required related amplifier pass-band. Where a resultant wave is the combination of two or more varieties of waves and one of these has a very steep wavefront, it is the portion of the resultant wave with the steep wavefront which dictates the frequency needs.

Phase or Time Delay in Amplifiers

We have shown that the nonsinusoidal wave is made up of a number of components of certain frequencies, each with a specific amplitude relative to the fundamental and each with a certain phase relationship to the fundamental.

In order that the amplified output be like the input wave, it is necessary that the amplitude and phase relationships between components be retained. This calls for specific performance characteristics on the part of the amplifier, especially those relating to phase shift or time delay. Ideally, the time delay of the different frequencies passed through the amplifier should be proportional to the frequency. The following will help make this point clear.

Phase delay or phase shift is the same as time delay. For example, if the frequency is 1,000 cps, 1 cycle or 360 electrical degrees requires an elapsed time of $^{1}/_{1000}$ or 0.001 second. If we say that the relative phase of this signal at one instant is 15° leading, this may be expressed in time as being $^{15}/_{360}$ or $^{1}/_{24}$ of a cycle or $^{1}/_{24} \times 0.001 = 0.0000416$ second or 41.6 microseconds. Naturally the elapsed time per degree differs with frequency so that the same condition at 100 cps would be the equivalent of 0.000416 second or 416 microseconds.

Relative to linear time delay or phase shift in an amplifier, the following illustrates how this condition is applied to the components of a complex wave and shows what is meant by phase shift proportional to frequency. Suppose that we are concerned with a wave which has a fundamental frequency of 1,000 cps and contains a second, a third, and a fourth harmonic. Relative amplitudes of the fundamental and its components will be disregarded, it being assumed that proper amplitude relationships will be maintained. Suppose further that at the instant of reference the fundamental frequency has a relative phase of 90° leading. This exists at the input of the amplifier and corresponds to a phase condition of $^{90}/_{360} \times 0.001$ or 0.00025 second. If the second harmonic is delayed proportionally to its frequency, it would be delayed by $2 \times 90°$ or 180°, so that the time delay is expressible as $^{180}/_{360} \times 0.0005 =$ 0.00025 second (the number 0.0005 is the time of 1 cycle of the 2,000-cps wave). If the third harmonic is delayed proportional to *its* frequency, it would be delayed by $3 \times 90°$ or 270° so that the time delay would be expressed as $^{270}/_{360} \times 0.000333 = 0.00025$ second. The fourth harmonic being delayed proportional to its frequency relative to the fundamental would be delayed by $4 \times 90°$ or 360° and the time delay or phase shift would be expressed by $^{360}/_{360}$ or $1 \times 0.00025 = 0.00025$ second. As can be seen, each component, as it passes through the amplifier, is delayed by exactly the same amount as the fundamental frequency, and this condition is expressed as linear, constant, or uniform phase shift or time delay.

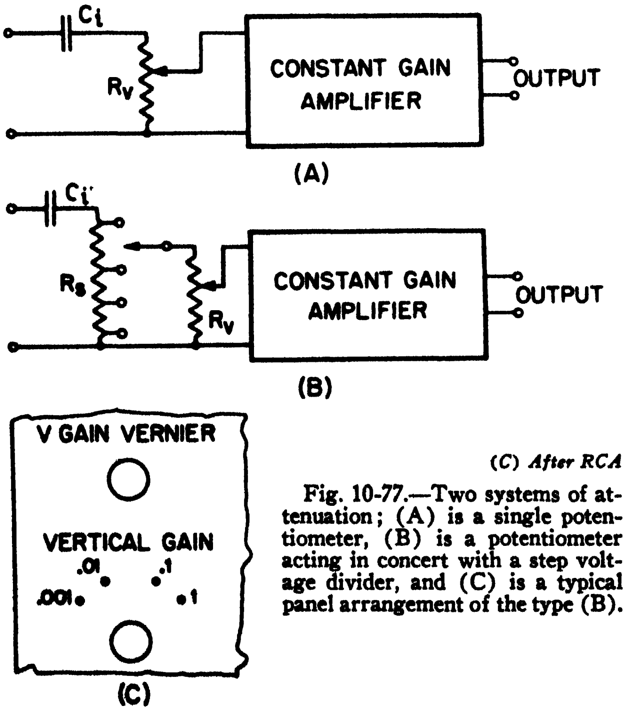
When this condition exists, the output signal is a true reproduction of the input signal, assuming that the frequency response of the amplifier is such as to afford substantial uniform amplification on all of the components for all of the fundamental frequencies. The selection of a fundamental and three harmonics does not imply that only three components can be treated in the aforementioned fashion. It should be understood that it is applicable to ten, twenty, or as many component frequencies as are present in the signal. Of course, the ideal conditions are not always attained, especially at the very high frequencies but this does not harm provided that the departure from the ideal is not too great. In other words we accept a slight amount of distortion. Once more here is an example of one of the reasons why the more elaborate oscilloscope amplifiers are more expensive to produce and, in the long run, cost more.

Linear time delay is a design characteristic determined by the designer of the amplifier, which means that the user has little choice in the matter. It is supposed that when an oscilloscope with certain rated characteristics is purchased, it affords those characteristics in its performance. The accomplishment of such phase characteristics over a band of frequencies does dictate certain specific contours in the over-all frequency response curve, so that it is in a way remotely possible by examining the response curve of the amplifier to see if the conditions of operation are such as to *expect* the proper performance relative to time delay. Such examination *does not* result in a quantitative answer; it is more of a qualitative result with implications, rather than definite information. It so happens that comparatively few oscilloscopes are described in part by an illustration of the frequency response curves, but in the event that such curves are available, the information about response curves which is given later in this chapter will no doubt prove helpful.

In view of the possibility that previous remarks about uniform amplification over the pass-band of either the horizontal or vertical amplifier may create the impression that nothing but the flatness of the response is important, it might be well to remark that uniform response accompanied by abrupt drop at both the high and low ends of the response curve is highly undesirable. It affords the required uniform amplification, but it will also introduce serious phase shift which will distort the output trace.

The Behavior of Vertical-Amplifier Attenuators

At first thought, one might consider the vertical-amplifier gain control a very simple part of the entire oscilloscope system since its duty is elementary. Such is not the case, however. In view of the fact that the various types of attenuators used in oscilloscopes are shown in detail in Chapter 22, we shall not attempt at this time to analyze individually the different attenuator circuits. Instead we will illustrate, in a general way, the effects they can have on the output waveform of the signal being passed through the vertical amplifier. Let us examine the attenuator from two viewpoints. One of these is its function as a means of controlling signal level. The other is its over-all effect on the frequency response of the amplifying system as the signal level is reduced. Each of these will be treated separately.



(C) *After RCA*

Fig. 10-77.—Two systems of attenuation; (A) is a single potentiometer, (B) is a potentiometer acting in concert with a step voltage divider, and (C) is a typical panel arrangement of the type (B).

The two kinds of attenuator systems have been mentioned earlier. This reference is an elaboration. One form of attenuator is a single continuous amplitude control. The other system is a combination of step control and continuous variation. These may be *symbolized* by a single potentiometer, and by a potentiometer acting in concert with a step voltage divider, as shown in Fig. 10-77A and B respectively; 10-77C represents a typical panel control arrangement of the type (B).

The need for step attenuators in addition to continuous variation arises from a number of conditions which must be satisfied and cannot be accomplished by means of a simple potentiometer. The range of signal voltages encountered in oscillography is very broad, yet a smooth control over relatively small increments of signal level is required. This is not possible when a single control which must be fixed in physical dimensions and still present a high impedance at the input side is used. If the resistance element of such a potentiometer could be made very large physically and be of a high resistance, it would satisfy these two needs, although they are neither the only, nor the most important, requirements.

The demand for faithful reproduction imposes certain frequency conditions as well as specific phase or time-delay conditions. These cannot be satisfied by the use of a single element potentiometer for reasons which will become evident soon, so that recourse must be made to the combination of a step attenuator and a continuously variable level control.

The Cathode-Follower Attenuator

The application of two-section attenuation systems is not as simple as in the symbolized circuit of Fig. 10-77B. While several methods are used, the one most frequently employed utilizes a cathode follower as an intermediate element between the step attenuator and

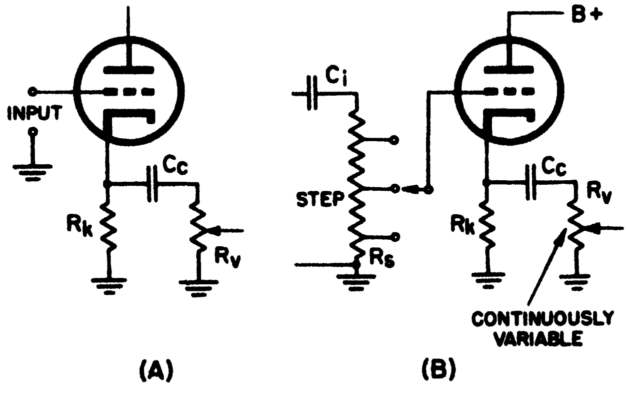


Fig. 10-78.—(A) is a variation of the attenuator system of Fig. 10-77(A) using a cathode follower before the potentiometer; (B) is a variation of Fig. 10-77(B) using a cathode follower between the step voltage divider and the potentiometer.

the continuously variable control. This is symbolized partly in Fig. 10-78A and fully in Fig. 10-78B. The reason for the two presentations is that the former may be used by itself over a limited range of voltages, although in most cases, the latter system also finds application.

Essentially, the cathode follower is a vacuum-tube impedance transformer; its input impedance is high and its output impedance is low and can be varied over an appreciable range. The gain A of such a stage is less than one, and is expressed as

$$A = \frac{R_k}{1/g_m + R_k} \qquad (10\text{-}9)$$

where R_k is the cathode load resistor and g_m is the mutual conductance of the tube. The gain of the cathode-follower can never be greater than unity because the quantity $1/g_m$ cannot become zero since g_m is finite. Thus the impedance transformation in this type of system invariably is accomplished by a loss in signal. Fortunately, the loss never is so great as to be a problem, whereas the transformation of impedance affords a definite advantage. The output impedance Z_L of the cathode follower is

$$Z_L = \frac{1}{g_m + 1/R_k} \qquad (10\text{-}10)$$

However, for our purposes, it is sufficient to view the cathode circuit as a low-impedance system which permits the application of a low-resistance potentiometer as the continuously variable gain control. The advantages of such a low-impedance device relative to frequency discrimination are discussed in a subsequent paragraph; a low-resistance control affords a smooth control of amplitude, and the presence of a high-impedance input is retained, with all of the advantages which it offers. These are discussed in more detail later in this chapter.

Capacitor C_o, which couples the potentiometer to the cathode resistor, is a blocking capacitor for d.c., and usually is a high value so as to present very little impedance to the passage of low-frequency currents. In the case of d-c amplifiers, this coupling capacitor is omitted, and the cathode resistor may be a potentiometer which serves as the gain control. The d-c voltage variations present in the cathode circuit are the signal voltages which the transferred to the succeeding stages.

In some systems which employ the step attenuator ahead of the first stage (the cathode follower) and the continuously variable element in the output of the stage, the lowest setting of the potentiometer does not reduce the signal to zero. This is done so as to necessitate the use of the step attenuator for full range of signal control, thus tending to minimize the possibility of overloading the input stage with an exceptionally strong signal. Because the variable control does not reduce the signal below a certain minimum, proper trace dimension can be accomplished only by adjusting the step control also. The attenuating systems symbolized in Fig. 10-77 include those used in almost all oscilloscopes. If we concern ourselves only with the basic oscilloscope, the attenuator shown in part (A) of that illustration would be the one most frequently encountered. The other varieties are representative of the more elaborate instruments.

When the multiple type, that is, the step and continuously variable arrangement is used, the initial setting of the step control should be that which affords the least attenuation consistent with a reasonably sized image; the continuously variable control should be wide open if possible. The reason for this may not be clear at the moment, but let it be said that from the viewpoint of frequency response to all types of nonsinusoidal waves, the best characteristics are generally attained when the gain controls are wide open, or at least as close to that condition as is consistent with the development of the required trace dimensions. The qualifying conditions follow.

The Frequency Effect of the Simple Potentiometer Control

Due to the presence of resistance and distributed capacitance in all electrical systems, the simple potentiometer arrangement shown in Fig. 10-77A is frequency sensitive. Distributed capacitance is to be found across the element as a whole and most certainly across that section which is delivering the voltage to the input circuit of the tube. Varying the position of the arm along the resistance element alters the constants of the input circuit, that is, the resistance and capacitance. The result is frequency distortion which is introduced on both sides of an optimum setting between the maximum and zero signal positions. This is shown in Fig. 10-79. These oscillograms illustrate the performance of a commercial oscilloscope when a 15,000-cps square wave was fed to the input. Although it might be said that too much was expected of the device because of its relatively limited frequency rating, the difference in pattern configuration using this high frequency illustrates most vividly the manner in which the simple potentiometer attenuator setting affects the reproduction. Oscillogram (A) was taken with the control near its zero point; oscillogram (B) for the optimum setting of the gain control, that is, the setting which resulted in a trace that most closely resembled the input

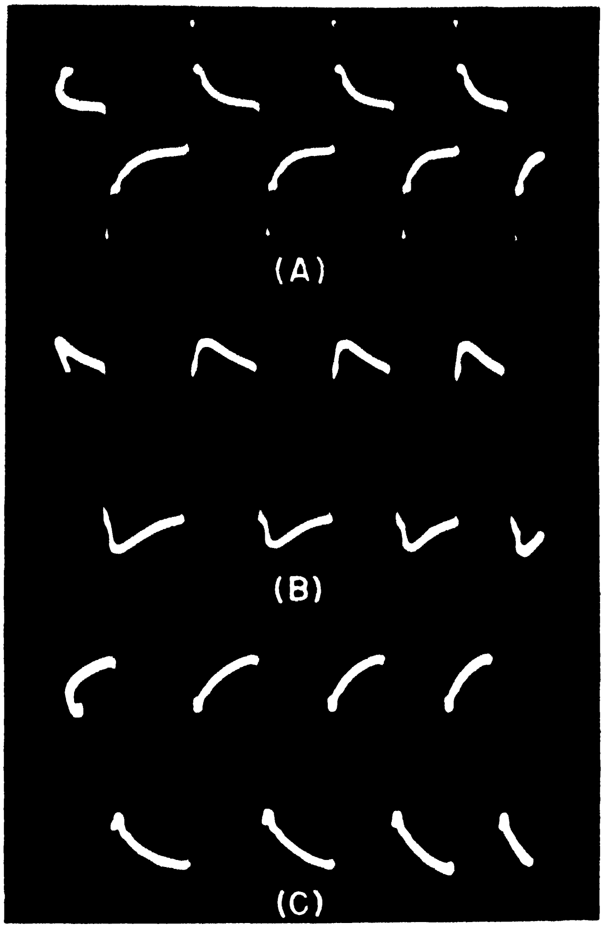


Fig. 10-79.—Distortion of a 15,000-cps square wave at various gain control settings; (A) results with the gain control near zero, (B) occurs at the optimum setting of the gain control, and (C) results with the gain control at maximum.

square wave; oscillogram (C) shows the trace when the gain control was wide open. The above is more frequently true with simple potentiometer-type attenuators of high value than with low-value units. This introduces another point of interest.

High- and Low-Resistance Potentiometers

It is always best if the input impedance of the vertical amplifier is high, that is, high resistance and low capacitance. The use of a high-resistance potentiometer tends to provide such an input condition. It is desirable in order to present the lowest possible load to the device or circuit which is connected to the input of the vertical amplifier. But, as we have mentioned, the high-resistance potentiometer attenuator gives rise to the condition stated. The alternative is a low-resistance unit. The difficulty with it, however, is that while it may give rise to less frequency distortion over its range of operation, its low resistance will load the circuit across which it is connected and, therefore, may influence the character of the output of that device.

As can be seen in Fig. 10-77, each of the attenuator systems is associated with an isolating capacitance, or a coupling capacitance. This unit is omitted in circuits designed for d-c amplification, but it is present in all a-c amplifiers, and where used contributes to the magnitude of the input impedance. If the resistance is high, the capacitance can be small and still permit good low-frequency response. But if the resistance is low, as would be the case when a minimum of frequency discrimination was desired by using a low-resistance potentiometer, a very high value of capacitance would be necessary. Since capacitance is a function of the dimension of the active surfaces, every condition which increases the dimension of electrical components tends to increase the distributed capacitance. Large capacitors and long leads connected to it tend to increase the shunt capacitance across the resistive section of the attenuator, and tend to defeat the purpose of the low resistance.

Before introducing another aspect of attenuators, it is imperative to associate the behavior of such attenuators with the usual interpretation of the over-all frequency-response curve of such units of a vertical amplifier. Although we have not as yet discussed the frequency-response curve of such units, we have indicated the desirability of uniform amplification over a range, and have indicated that vertical amplifiers in oscilloscopes do bear certain frequency bandwidth ratings. When the interpretation of these bandwidths becomes a concern, or when one is evaluating the utility of a uniform amplification rating, it cannot be done completely without taking into account the action of the attenuator.

The Attenuator and Amplifier Frequency Response

A frequency-sensitive or discriminatory attenuator system defeats the purpose of a uniform gain rating over a certain bandwidth. If the rating is predicated on sine-wave amplification, and it is so indicated, all well and good, but an interpretation of a sine-wave rating in terms of nonsinusoidal amplification is very apt to give misleading results. As we said before, simple attenuation of the high frequencies, or attenuation of low frequencies during sine-wave amplification, requires nothing more than correction of the gain versus frequency factor. In fact, if the rating is made with the attenuator wide open, the over-all response figures include the behavior of the attenuator. But if the ampli-

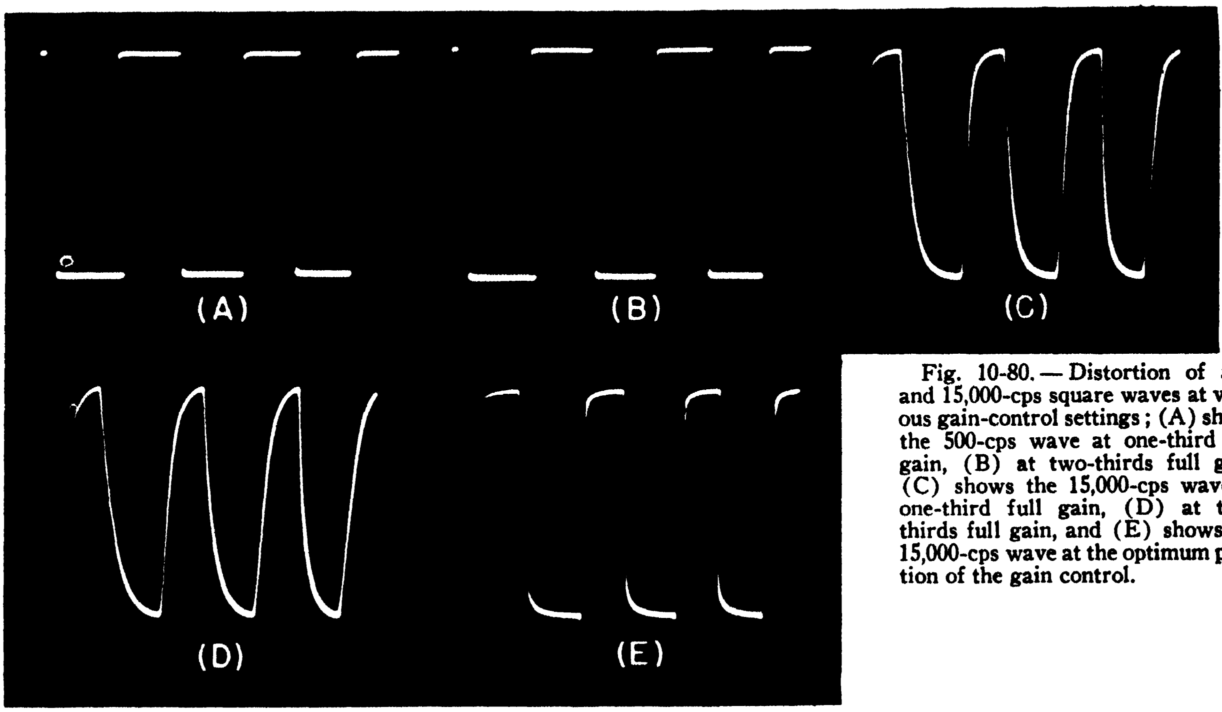


Fig. 10-80. — Distortion of 500- and 15,000-cps square waves at various gain-control settings; (A) shows the 500-cps wave at one-third full gain, (B) at two-thirds full gain, (C) shows the 15,000-cps wave at one-third full gain, (D) at two-thirds full gain, and (E) shows the 15,000-cps wave at the optimum position of the gain control.

fier is to be used for the examination of steep front waves or for their reproduction, simple numerical correction for differences in gain versus frequency at different frequencies means nothing. It then becomes necessary to limit the range of fundamental frequencies of such waves to within boundaries which are not only within the over-all bandwidth of the amplifier, but also within the range where the action of the simple attenuator is not too disturbing. This is a serious restriction for those who have occasion to use oscilloscopes for the observation of square pulses and other nonsinusoidal waves.

Exactly what we mean is shown by the oscillograms of Fig. 10-80. These are representations of two different square waves, one of 500 cps and one of 15,000 cps. Each was fed into the vertical amplifier of an oscilloscope which would be placed in the category of a general purpose device; it is perhaps not the simplest of this class, but is one which makes use of an uncompensated attenuator consisting of a high-resistance potentiometer.

Oscillograms (A) and (B) show the 500-cps square wave as reproduced with one-third full gain and two-thirds full gain positions, respectively, of the attenuator control. A slightly increased curvature at the leading edges is visible in (B) but it is of no importance; the representation would be adequate for all purposes.

Using the same two settings, but applying a 15,000-cps square wave, the results are shown in oscillograms (C) and (D). Definite loss of high frequencies is indicated, worse in (D) than in (C). To show that the condition is a function of the setting of the gain control, rather than the amplifier response, oscillogram (E) illustrates the reproduction of the same 15,000-cps square wave at that attenuator setting which afforded the closest approach to the shape of the input signal. Inasmuch as oscillograms (C), (D), and (E) differ only in the setting of the attenuator, it is easy to see that with an uncompensated attenuator one cannot determine the performance of the oscilloscope by examining the amplifier bandwidth only.

If this oscilloscope were examined from the viewpoint of the frequency range of square waves which could be fed into it without any effect due to the attenuator setting, the range of input frequencies of nonsinusoidal waves would be appreciably less than 15,000 cps, for, as indicated by Fig. 10-80E, its acceptance as a 15,000-cps square wave is questionable.

In order to evaluate the vertical amplifiers, the performance ratings must be read carefully. Since the characteristics of the attenuator are frequently stated, erroneous conclusions due to failure to properly interpret the equipment specifications are burdens which the user must bear. It so happens that up to the time of this writing, standards concerning oscilloscope specifications do not exist, so that an understanding of the differences between devices and the contributions of the different sections or parts of sections is necessary for the exercise of proper judgment when purchasing an oscilloscope.

Frequency-Compensated Attenuators

The action of uncompensated attenuators clearly indicates the need for devices which will not cause frequency distortion, yet will permit control of the signal level. They exist and are known as *frequency-compensated attenuators.* They are of the dual unit variety and are so used with the continuously variable element that the latter becomes a low-impedance device. This minimizes the effect of associated distributed capacitance, yet allows a high input impedance to the vertical amplifier as a whole. In practice, the frequency compensation is applied to the step attenuator at the input of the amplifier, whereas the continuously variable control is located at the output of the first stage, or at some point in the amplifier, where its manipulation will control the signal level consistent with minimum distortion. Examples of these are numerous in Chapter 22.

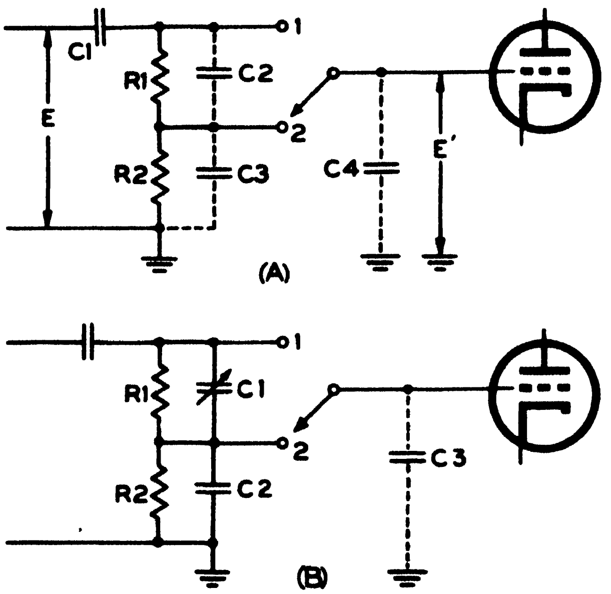


Fig. 10-81.—Schematics of step attenuators; (A) uncomsated, (B) compensated.

To follow the manner in which frequency compensation is accomplished, consider the circuit shown in Fig. 10-81A. This is a simple two-step attenuator at the input of a tube. The toggle switch is one example of this mechanism. It consists of a capacitor *C1* in series with a fixed resistor which is tapped at a point so that the voltage between ground and tap *2* is one-tenth of the voltage between ground and tap *1.* Associated with each section of the resistor and with the resistor as a whole, also at the input of the tube, are different amounts of distributed capacitance. These are shown in dotted lines.

For the initial discussion, we will neglect the presence of the distributed capacitance and consider only *C1, R1,* and *R2;* these three elements form a voltage divider. With the switch set on tap *1,* the two sections of the divider are *C1* and the resistance (*R1* + *R2*). Regardless of where the switch is set, only a portion of the voltage *E* can reach the input of the tube because the reactance of the capacitor acts as a voltage-dropping element. Whether this drop across *C1* is great or small depends upon the frequency of the voltage. The portion of *E* reaching the tube input depends on the relative values of reactance X_{C1} and resistance (*R1* + *R2*). With the switch set at position *1,* the equation for *E'*, the voltage at the tube input, is

$$E' = \frac{R1 + R2}{\sqrt{(R1 + R2)^2 + X^2_{C1}}} \times E \qquad (10\text{-}11)$$

and if the switch is set at position *2,* the equation becomes

$$E' = \frac{R2}{\sqrt{(R1 + R2)^2 + X^2_{C1}}} \times E. \qquad (10\text{-}12)$$

Regardless of the frequency, the higher the reactance of *C1,* the greater will be the drop in voltage across it, but the higher the value of the resistance relative to X_{C1}, the less the voltage drop across the capacitive reactance relative to the voltage drop across the resistive portion of the system. If we imagine the frequency very low, so that X_{C1} is several times higher than *R1* + *R2,* comparatively little of the voltage *E* will appear across *R1* + *R2,* whereas if X_{C1} is very low in comparison with *R1* + *R2,* most of the voltage will appear across the resistance. If we take license with the actual frequency and assume X_{C1} to be a very small fraction of *R1* + *R2,* then we can say that the presence of *C1* has very little effect on the signal level, although it does act as a blocking capacitor for direct currents.

It is evident that if a nonsinusoidal wave containing a wide range of harmonics is fed into such a system, the amplitude relationship of the harmonics will not be retained at the output of the divider system. Of course, it is possible to visualize a value of *C1* which will make X_{C1} extremely small relative to *R1* + *R2* at the lowest possible frequency, so that the different degree of attenuation which will take place due to changes in X_{C1} will be negligible. This is possible but not practical because of cost and size, and because it is not the entire solution.

Let us examine the system from another viewpoint. The capacitor *C1* and the resistance element comprise an *R-C* system with a certain time constant. It is possible to establish such values for C and R that the resultant time constant is sufficient to satisfactorily accommodate voltages of the lowest frequency which may be fed into the system. If this is done by the use of a very high value of *C1*, the increase in cost, dimensions, and the distributed capacitance may be undesirable. If, on the other hand, a compromise is made and the resistance is made large, it may be too high relative to the permissible resistance in the grid circuit of the tube. The optimum value of time constant selected for any such system, as reflected in the constants of the elements, must of necessity be a compromise, because it is virtually impossible to establish a single system which will perform uniformly in all respects at all frequencies. A certain amount of time delay is expected and accepted; invariably this takes place over the lower end of the frequency range.

Now, if we take cognizance of the distributed capacitance which we have shown as *C2, C3,* and *C4* in Fig. 10-81A, the action of the attenuator over a range of frequencies is still further modified. With these capacitances active in the system, and the switch set at tap *1*, the magnitude of the voltage fed to the input of the tube now is influenced by the distributed capacitance as well as by series capacitor *C1*. These values of distributed capacitance will have much greater effect at the high frequencies than at the low frequencies.

Since these shunt reactances are high at the low frequencies, the division of voltage between X_{C1} and $R1 + R2$ will be affected very little whereas when the frequency is high, the division of voltage no longer is determined by the values of *R1* and *R2*, but rather by combined effect of the reactance of the distributed capacitances and the resistive elements which they shunt. In other words, the entire attenuator system becomes frequency sensitive; changing the tap position may afford the proper voltage division over a limited frequency range, but not over the entire range. Under the circumstances, changing the tap positions when the input voltage is of nonsinusoidal character will treat the different components differently and tend to distort the reproduction. The presence of these capacitances will also modify the time constant of the system as the tap positions are changed, and introduce varying time delays. So, both amplitude attenuation and phase shift are the problems of uncompensated step attenuators, just as they are the problems of the simple potentiometer type of dividers.

Compensation in a step attenuator is accomplished by the use of additional shunt capacitances; it is customary to make one or more of these capacitances variable to permit adjustment for variations in circuit capacitance. A simple arrangement is shown in Fig. 10-81B. Capacitors *C1* and *C2* are the added compensating capacitors, and are considered to include the switch and wiring capacitances across the taps. *C3* represents the tube and additional circuit capacitances. Neglecting the series blocking capacitor, the voltage at switch position *2*, as determined *only* by the resistive divider, will be proportional to $R2/(R1 + R2)$. In order that the capacitive shunting reactances also form the same voltage-divider ratio, the reactance of *C2* and *C3* in parallel, divided by the reactance of *C1* in series with the reactance of *C2* and *C3* in parallel, must be equal to $R2/(R1 + R2)$. When this is accomplished, the attenuator is relatively independent of frequency over a very wide range of frequencies.

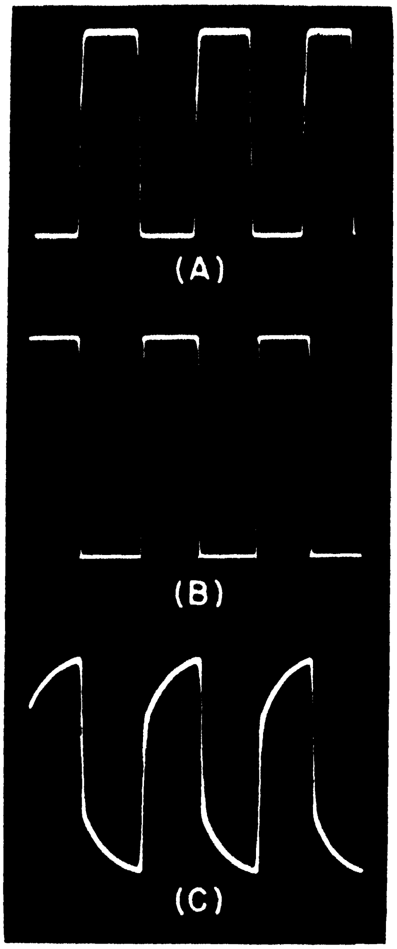


Fig. 10-82.—Illustrating the effect of a properly compensated attenuator, (A) the trace of a 15,000-cps square wave fed directly to the deflection plates, (B) the pattern of the same square wave at a point on a properly adjusted attenuator, and (C) the result of improper adjustment of the attenuator.

The presence of compensation in a step attenuator does not necessarily mean that all variation in the constants of a system has been removed. All it means is that a suitable compromise has been accomplished

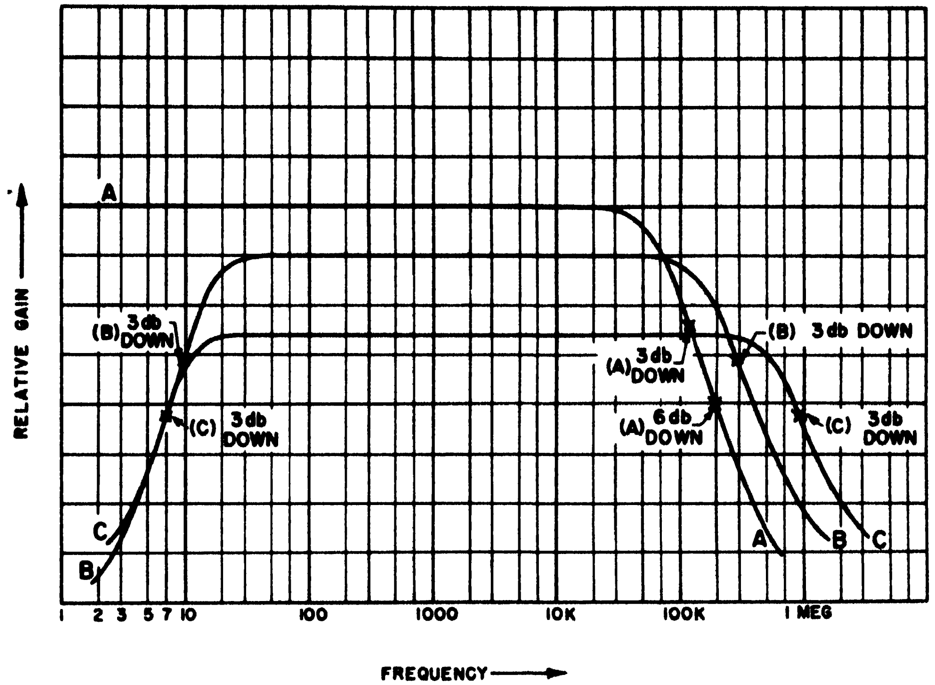


Fig. 10-83.—Typical frequency response curves.

which minimizes the distortion in the reproduction to a point where it does not impair the utility of the oscilloscope. Such compensation is a very accurate procedure and it is entirely possible that, after a period of use, some changes will take place in either the adjustment of the trimmers or in the constants of the circuit. When this happens, it will influence the performance of the oscilloscope when the defective switch step is used. An example of this is shown in Fig. 10-82. Oscillogram (A) is the trace of a 15,000-cps square wave when it was fed directly to the deflection plates of the cathode-ray tube. Oscillogram (B) shows the pattern of the same voltage at one point on a step attenuator which was properly adjusted, while (C) shows the trace when the components of the same step on the attenuator were incorrectly adjusted. The effect of the greatly increased attenuation at the high frequencies is very evident.

Frequency-Response Curves

So many references have been made to frequency response and bandwidth that a short discussion of frequency-response curves of the vertical amplifier cannot be avoided. While it has not been usual commercial practice to illustrate the performance of these amplifiers in this way, some manufacturers do so and the method is gaining favor. Therefore, a few brief remarks of explanatory nature may make these graphs more understandable, especially from the viewpoint of use.

A frequency-response curve is a graphical presentation of the manner in which an amplifier performs relative to frequency, that is, over the frequency rating of the system. In Fig. 10-83 are shown three response curves. The ordinate or vertical axis indicates the level of the response or the amount of amplification which the amplifier affords. The frequency scale is along the abscissa or horizontal axis.

The usual method of indicating the amount of amplification is in terms of relative gain. The amount of amplification available at any one frequency relative to that available at another is indicated by the heights of the response curve. The three curves in Fig. 10-83 convey the information that each of the amplifiers affords a different amount of amplification over the same frequency range.

The complete range over which the amplifier will respond is also shown by the curves. This is indicated by the span of each response curve relative to the frequency axis. Curve *A* indicates that the amplifier it represents affords amplification down to zero frequency (d-c),[1] which is below the limit of amplifiers *B* or *C*. However, at the high end of the scale, amplifier *A* (with the greatest gain of the three) cuts off sooner than either amplifier *B* or *C*.

On the other hand, amplifiers *B* and *C*, each with different bandwidths, start at substantially the same low frequency. This is not out of the ordinary because it is a function of design. For that matter, the relationships between gain and bandwidth as depicted in this curve are along the normal lines of design, that is, the greater the bandwidth of an amplifier, usually the less is its gain. This should not be interpreted as a fixed rule which will apply to any two commercial devices because individual designs may differ, and an amplifier with wider response may also afford greater gain than another commercial amplifier. The rule stated is one which applies to a given amplifier during design.

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[1] The curve shows the amplifier response down to only one cycle, since there is no zero on a logarithmic scale. In practice, this is the typical manner in which the low-frequency end of a d-c amplifier response curve is drawn.

An important aspect of such frequency-response curves is that which relates to the types of signals which are fed into the amplifier. The broad interpretation of the amplifier response as shown in the response curve applies only to the amplification of sine waves. As indicated, these amplifiers are flat (uniform amplification) over a range of frequencies; at both the upper and lower limits, the gain falls off, finally reaching zero. However, even those regions which afford reduced amounts of gain are useful for amplifying purposes, provided that the signals are sinusoidal, and that the amount of reduction in gain is known.

Both of these statements require some qualification. If we have sine-wave voltages in mind, and one of the functions of the oscilloscope is to measure voltage levels by means of the height of the trace on the screen, then it becomes important to know the exact characteristics of the amplifier. It can readily be seen that a frequency which is within the zone of operation where the amplifier is flat will receive one level of amplification, whereas a frequency within the zones of reduced gain will be subject to a reduced amount of amplification. To place these two on par will require the application of a correction factor equal to the reduction in gain of the amplifier.

As stated previously, sine-wave frequencies which may be applied to such amplifiers extend over the entire range indicated by the response curve. This is interesting because the response curve includes a frequency range which exceeds the rated pass-band of the amplifier, as mentioned in the specifications. This rating restricts the response to that range of frequencies which is amplified to within 3 db of the level which represents the "flat" part of the amplifier. According to this practice of rating, both the high- and low-frequency limits of the amplifier are those frequencies which receive not less than 70.7 per cent of the maximum amount of gain available in the amplifier.

Referring to curve *A* of Fig. 10-83, it can be seen that the amplifier has no cutoff at its low end, but does have an upper limit at slightly above 100 kc, the highest frequency of which receives 70.7 per cent of the amplification over the flat portion of the curve. By no means is 100 kc the actual limit of response in this amplifier; the response curve extends beyond the 3-db point, but the extended range is not included in the rating of the response. The frequency 6 db down is 200 kc.

If the frequency rating included the total range of frequencies which can be amplified by the device, although to varying degree, it would be misleading when interpreted in terms of square-wave or other nonsinusoidal frequencies which were within the capabilities of the device. You will recall that the development of an acceptable version of a nonsinusoidal wave demands more than the mere presence of a certain number of harmonic components; the reproduction of such a wave calls for *uniform* amplification of these components and for a linear phase relationship.

Limits of Bandwidth

The limited frequency-response rating stems from the distortion which is introduced by the presence of components when they are substantial in their importance to the reproduction, but do not receive adequate amplification with respect to the lower-frequency components. For example, the amplifier represented by curve *A* with a rated 3-db cutoff of 100 kc would be suitable for the amplification of square waves with a fundamental of 2,500 cps, assuming the presence of twenty odd harmonics, or 5,000 cps, assuming the presence of ten odd harmonics. If we used the 6-db point as the upper limit of the range, and assumed the presence of twenty odd harmonics, the highest square wave acceptable for reasonable reproduction would be more than 5,000 cps, whereas for ten odd harmonics, it would be 10,000 cps.

Referring once more to Fig. 10-83, curve *B* might be that for an amplifier which would be rated as having a response of from 10 cps to about 300 kc. Both limits are indicated as the 3-db points on the curve. Curve *C*, on the other hand, is for an amplifier having a frequency response of from 7 cps to 900 kc.

Some manufacturers rate the amplifiers in their oscilloscopes somewhat differently. They are a bit more severe and consider the bandwidth as being between the 2-db points, which are equal to 79 per cent of the maximum gain, or about 20 per cent down. It must be mentioned, however, that such references simply state the upper-frequency limit without implying any upper limit of square-wave frequency which may be fed into the amplifier and which will result in acceptable reproduction.

Another specification which is frequently found, and which is to be used at the discretion of the user, is the highest frequency which suffers 50 per cent loss in amplification, or is 6 db down.[2] Correlating these methods of indicating response with the curves of Fig. 10-83, it stands to reason that the less the allowed loss in amplification, the narrower will be the rated pass-band.

We cannot stress too strongly the difficulty of evaluating oscilloscope capabilities from specified amplifier bandwidths. No matter how definitely mathematical computations may indicate that a certain number of harmonics are necessary for an acceptable reproduction, the practical applications of oscilloscopes frequently show otherwise. The results indicate that fewer harmonic components frequently are satisfactory. This is illustrated in Fig. 10-84. Here we show a number of oscillograms of a 20,000-cps square wave. The amplitude of the wave was maintained relatively constant on the screen by manipulation of the output control of the square-wave generator for progressive positions of the simple potentiometer-type attenuator.

By all sound reasoning, this oscilloscope, which has a rated bandwidth of 200 kc, 2 db down, and a rated bandwidth of somewhat more than 300 kc, 6 db down, should not be suitable for the reproduction of square waves of this frequency. Nevertheless, here is testimony that the development of a reasonable square wave (C) and (D) was a function of the attenuator position. If we could assume the removal of the attenuator, thereby permitting the output trace to be a function of the amplifier characteristics only, the chances are that this comparatively low pass-band is capable of producing an acceptable square wave, even though only about seven odd harmonics are present within the 2-db points. This seems to indicate that amplifier bandwidth ratings may be made more liberal if the information being sought is qualitative, and that the contributions made by the frequencies which suffer loss in amplification beyond the 3-db point may not be too harmful in many applications of the oscilloscope. This is, however, strictly a practical approach.

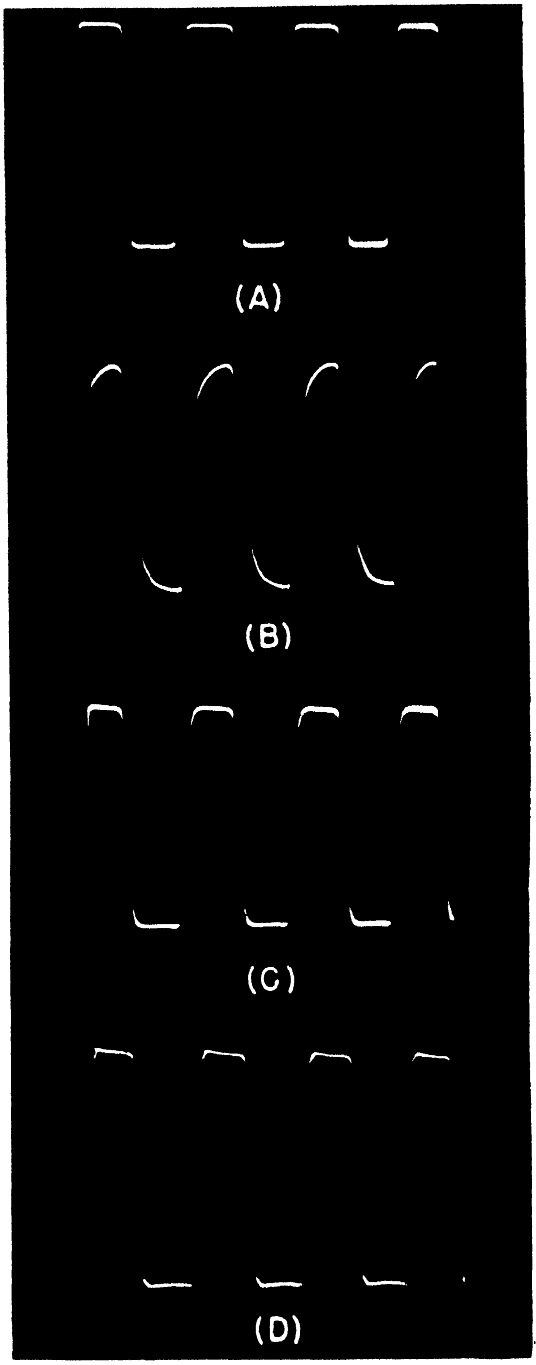

Fig. 10-84.—(A) 20,000-cps square-wave input, (B) trace with improperly adjusted attenuator, (C) and (D) attenuator adjusted to produce reasonable reproductions of the square wave.

Basis of Wide-Band Response

The over-all response curve of an oscilloscope amplifier is a composite of the individual responses of the different stages which comprise the amplifier. If the individual stages were checked separately, the response curves would differ markedly from the total of them all. We might go so far as to say that the stages are individually *tailored* so as to produce a definite resultant, the extent of such individual treatment being more critical as the width of the pass-band is greater.

It is not within our province to examine the organization of amplifiers in this respect, but we must say that the design and construction of such a system is

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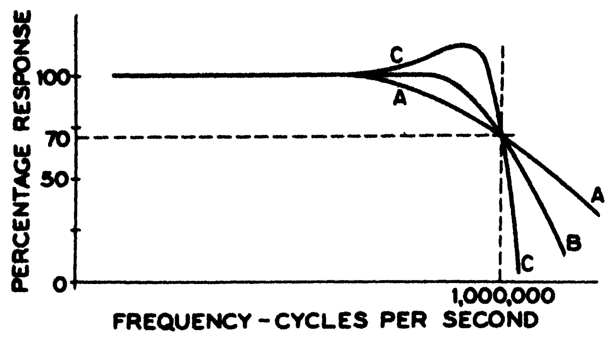
[2] 10% down approximates −1 db
20% down approximates −2 db
30% down approximates −3 db
50% down approximates −6 db

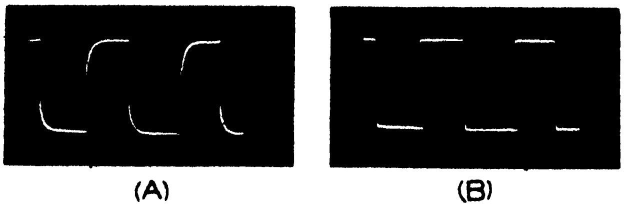


After Du Mont Labs.

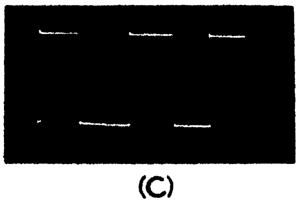
Fig. 10-85.—Frequency response curves with different degrees of compensation, (A) insufficient, (B) proper, and (C) excessive.

very critical and changes in components or *tubes cannot be made indiscriminately.* (Compensating for nonlinearity in time-base oscillators is explained in Chapter 9.) Lead dress is important and even a slight alteration in constants is enough to modify the shape of the response curve, and materially alter the reproduction of the signal fed into the amplifier.

The means of attaining wide response is the creation of quasi-resonant states in the individual stages, each of which may be tuned to a different frequency within the over-all desired band. Each stage then contributes a certain amount of the gain. The adjustments are critical. The effect of different degrees of compensation as well as the effect of insufficient compensation is shown in Fig. 10-85. Three amplifier response curves are illustrated, each of which has the some nominal bandwidth. Curve *A* slopes gradually at the higher frequencies and is representative of the amplifier without high-frequency compensation, that is, an amplifier in which no attempt has been made to build up the region of higher-frequency response. Curve *B* is flat to a higher frequency because compensation has been applied. (Just how this is done will be shown soon.) The sign of such compensation is the more abrupt decrease in gain, although a too rapid fall in gain can be troublesome. Actually, it is difficult to determine the character of compensation from the response curve, unless it is extreme to the point where the taper at the high end is almost a straight line downward. Such a drop-off is not desired. Curve *C* shows a very steep drop, but also a fairly sharp rise in response at the high end. This is an example of overcompensation and the two conditions usually accompany each other. It might be said that unlike tuned r-f band-pass amplifiers, where very steep sides are desired, oscilloscope amplifiers demand more gradual slopes in order to minimize phase distortion. The action of three such amplifiers on a 100-kc square wave is shown in Fig. 10-86, which indicates that the most favorable type of response curve is that of Fig. 10-86B.



(A) (B)



(C)

Fig. 10-86.—Traces of a 100-kc square wave after passing through amplifiers with responses corresponding to Fig. 10-85(A), (B), and (C), respectively.

Courtesy Du Mont Labs.

The overshoot in Fig. 10-86C indicated by the narrow pip at the top of the leading edge of the wave is evidence of excessive high-frequency amplification or overcompensation. Expansion of this wave so as to show the character of the flat portion of the wave more clearly would produce a trace like that shown in Fig. 10-87. This type of amplifier operation is frequently called "ringing" and may develop if some change takes place in the condition of the frequency-compensating components. If a state of overcompensation exists in the oscilloscope amplifier, it will, as a rule, indicate its presence as the fundamental frequency of the square-wave testing signal is increased.

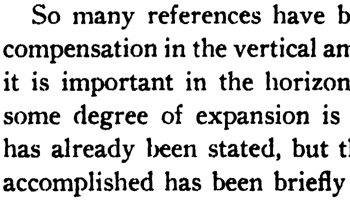


Fig. 10-87.—Enlargement of Fig. 10-86(C).

Frequency-Compensation Methods

So many references have been made to frequency compensation in the vertical amplifier (for that matter, it is important in the horizontal amplifier, too) that some degree of expansion is necessary. Its function has already been stated, but the means whereby it is accomplished has been briefly referred to. In view of its importance in the proper behavior of the oscilloscope, even though it is taken care of in the design, and since all signs point to even greater stress being placed on the pass-band of the amplifiers in oscilloscopes, we feel that the subject deserves its full measure of attention. Accordingly, at the end of this chapter we have included an appendix containing an analysis of the subject giving the necessary design information which might prove helpful to any one interested in the design of an amplifier, or who might desire to explore the subject more than just casually.

The Advantages of Wide-Bandwidth Amplifiers

It would seem from the foregoing that maximum emphasis was being placed on the presence of the fewest possible harmonics in a reproduced wave, consequently on the amplifier with the narrowest bandwidth. We have no particular reason for taking a stand of this kind, in fact, the purpose of this and the subsequent paragraphs, that are related to the subject, is to correct any possible misunderstanding. We are in no way favoring a 300-kc bandwidth oscilloscope in preference to one with wider bandwidth. We do, however, say that a minimum of seven odd harmonics is sufficient to afford a reasonable facsimile of a square wave. Moreover, such reproduction is adequate for numerous television maintenance operations.

Now for the other side of the subject, let us reiterate that the greater the bandwidth is, the better generally. For example, let us assume that a very-high-frequency parasitic oscillation is present in an amplifier, or for example, an overcompensated amplifier is to be checked. Both of these conditions may or may not be associated with television maintenance, but the possibility of similar conditions being present in a television system is very great. Their determination by means of an oscilloscope with relatively narrow bandwidth is very remote, if possible at all. In other words, a 700-kc transient would not be discernible with a 200- to 300-kc bandwidth amplifier.

The wide-band oscilloscope offers many operating capabilities that are beyond the province of the medium-bandwidth and low-band instruments. Every activity can neither afford, nor does it require the same kind of oscilloscopes. The selection of an instrument depends, therefore, upon what is needed rather than the electrical superiority of one over another. Any recommendations implied herein use the minimum capabilities necessary for certain duties, but the electrical limitations which accompany the compliance with such requirements also must be recognized.

CIRCUIT FEATURES OF VERTICAL AMPLIFIERS

The circuits used for the vertical-deflection amplifiers may be the same in different kinds of oscilloscopes, and may differ quite substantially, even among the basic varieties of devices. Many instruments make use of resistance-capacitance of *R-C* coupled systems, others employ direct coupling, but the manner in which the circuit elements are treated, the number of tubes used per stage, the use of single-ended and push-pull stages, the varied forms of phase inversion, and, finally, the varied degrees of frequency compensation represent the differences between oscilloscopes of different categories, functional capabilities, and certainly price class.

A-C and D-C Amplifiers

Two fundamental varieties of circuits are used for interstage coupling between tubes in vertical amplifiers. These are the *R-C* system previously mentioned, which is also known as the *R-C* coupled amplifier, and the direct-coupled amplifier which is frequently called the d-c amplifier. The first of these is shown in simplified form in Fig. 10-88.

In the *R-C* coupled system of Fig. 10-88, three important circuit elements must be taken into account. These are the plate-load resistance *R1*, the coupling capacitor *C* (sometimes known as the blocking capacitor), and the grid-leak resistance *R2*. In view of the location of *C*, the grid of *V2* is isolated from the d-c plate voltage of *V1*. In this respect, it is a blocking capacitor. Let us now examine how a signal is transferred from *V1* to *V2*.

Assuming a quiescent condition in both tubes, that is, no signal, the effective voltage at the plate of *V1* is steady and of a value equal to the B+ voltage minus the drop across *R1* due to the current I_b which is present in the plate circuit. If we can imagine an uncharged capacitor *C* in series with resistor *R2* being placed into the circuit as shown in Fig. 10-88, a charging current i_c will flow through *R1-C-R2* until *C* becomes charged to a voltage corresponding to the plate voltage of *V1*. This is momentary action but during the time the charge in *C* is *changing*, the charging current through *R2* will produce a voltage drop i_cR2 which will be applied to the grid-cathode circuit of *V2* as a momentary signal. Its value, direction, and duration will depend on those of i_c.

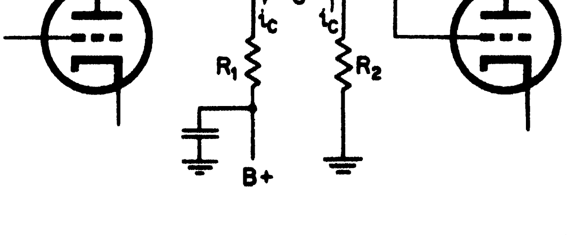


Fig. 10-88.—Simplified schematic of *R-C* coupled amplifiers.

When capacitor *C* becomes charged to a value corresponding to the original charging voltage, movement of charges through the *R1-C-R2* circuit ceases. The voltage drop across *R2* disappears, all variables in the circuit have become constant and the grid in *V2* experiences no effects from the previous momentary signal voltage; the potential of that grid becomes whatever is dictated by the grid bias applied to that tube, just as if there had been no movement of charges in *R1-C-R2*.

Let us now apply a signal to the control grid of *V1* so that a change takes place in the plate current of that tube. This will change the voltage drop across *R1*, which element is fixed in ohmic value. Since the voltage drop across this resistor determined the original charging voltage for capacitor *C*, a *change* in this voltage will result in a new charging current, that amount necessary to make the voltage across *C* equal to the new value of plate voltage or charging voltage.

The path of this *charging* current is through *R1-C-R2*, consequently, a momentary voltage (signal) will again appear across *R2*, which is the same as saying that it will appear across the grid-cathode circuit of *V2*. The magnitude of this signal input to *V2* is determined by the value of i_b, or the momentary *change* in plate voltage of *V1*, and while it lasts, it changes the potential of the *V2* grid relative to its cathode so that a signal has again been transferred from the grid of *V1* to the grid of *V2*.

Now the interesting aspect of this action is that a signal appears across *R2* only during *changes* in the plate voltage of *V1*. We can restate this by saying that changing plate current through *R1* causes a changing value of charging voltage to be active relative to the circuit of *CR2*. During these times the grid of *V2* is likewise changing in potential relative to its cathode in conformance with the changes in the voltage i_c *R2*, and the capacitor *C* is behaving like a coupling capacitor.

The direction of the charging current through *R1-R2* and into and out of *C* depends on the polarity of the changes on plate voltage of *V1*, and determines whether the potential of the *V2* grid will momentarily rise or fall (will become less negative or more negative) relative to its static, no-signal value as determined by the applied grid bias. The instant that charging current ceases to flow into or out of *C*, therefore, through *R2*, the potential of the *V2* grid returns to its original no-signal value just as if nothing had happened previously. This is illustrated in Fig. 10-89A and B.

Therein are shown the conditions at the grid of *V2* with steady plate voltage on *V1* and a grid bias of −4 volts on the grid of *V2*, and change in grid potential at *V2* when the plate voltage of *V1* is varying as a sine wave, increasing and decreasing around the steady no-signal value. It is evident that when the plate voltage of *V1* is constant, the no-signal condition prevails at *V2*. Also that each time the plate voltage of *V1* returns to its no-signal steady value which corresponds to zero change, the grid potential of *V2* returns to its steady no-signal bias value. In between, the grid of *V2* goes more negative and less negative according to the value and polarity of i_c *R2* with respect to the fixed bias of −4 volts.

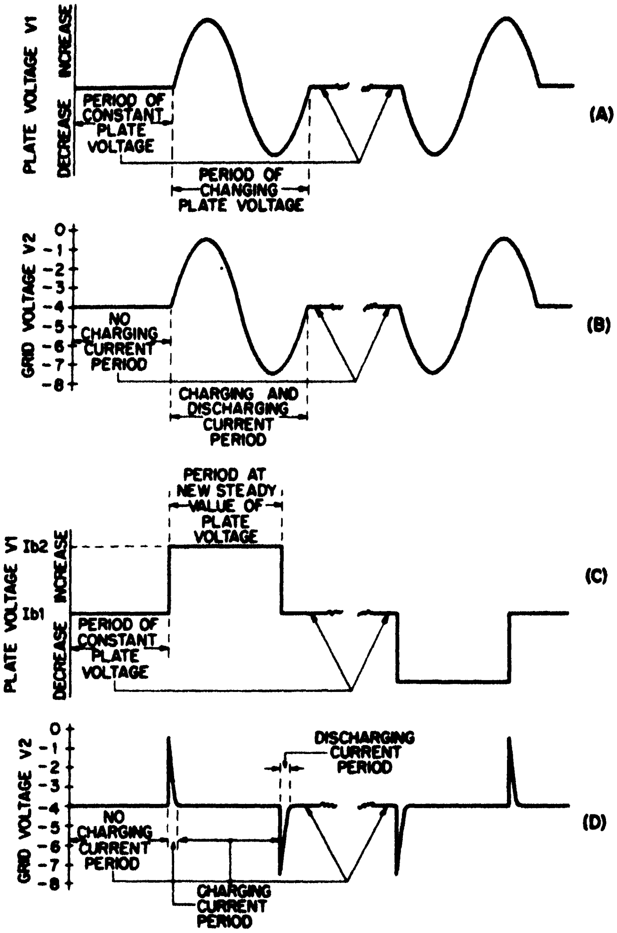


Fig. 10-89.—Illustrating the relation between the driving plate voltage and the driven grid voltage in *R-C* coupled amplifiers; (A) and (B) for sine-wave signals; (C) and (D) for square-wave signals.

Let us now imagine a change in bias on *V1* so that the value of plate voltage changes from E_{b1} to a new value E_{b2} and remains constant at the new value for a while, and then, the bias is restored to its original level which naturally returns the plate voltage to its original value E_{b1}. This action of the plate voltage is shown in Fig. 10-89C. Visibly, the change in grid bias is unidirectional since the change in plate voltage likewise is unidirectional.

What happens in the coupling system *R1-C-R2* and at the grid of *V2*? Let us examine Fig. 10-89D as well as C. During the no-signal condition in *V1*, the plate voltage is steady and the grid of *V2* is at the bias potential of −4 volts. Then there occurs a sudden change in plate voltage at *V1*, which means:

1. A new value of charging voltage
2. A sudden rush of charging current through *R1-C-R2* in the attempt to change the potential of *C* to conform with the new charging voltage
3. The appearance of a momentary voltage drop across *R2*, therefore, a sudden change in the potential of the *V2* grid
4. When the momentary change in plate voltage of *V1* has been accomplished and it is at its new value, the coupling capacitor *C* also has been charged to the higher voltage. There it stays with its charge and the charging current through *R2* (also *R1*) decreases to zero — and concurrently with that, the grid of *V2* returns to the *original* value which prevailed before anything happened in the plate circuit of *V1*.

Sometime later the plate voltage of *V1* is restored to its original value. The change in plate voltage becomes a new value of charging voltage and the capacitor discharges some of the energy it has stored. The discharge current is a momentary rush of charges through *R2-R1*, and a momentary voltage again appears across *R2*, but now in the opposite direction to what it was before, and the grid of *V2* assumes a potential for an instant. After the coupling capacitor has been discharged sufficiently so that it is at the voltage of the plate of *V1*, the discharge current ceases, and the potential of *V2* grid assumes its normal value.

Summarizing the contents of Fig. 10-89, it is easy to see that the transfer of a signal through an *R-C* amplifier demands a continuous change in voltage conditions in the coupling system, a steady-state condition, such as a d-c change will be accomplished only temporarily, namely, during the moments of transition from one state to the other. This is shown in Fig. 10-89C and D, where the change in value of plate voltage from one level to the other is a change in steady-state conditions. Only when there occurs a continuous change in state as shown in Fig. 10-89A and B, does there take place a continuous transfer of energy.

Now if we consider a continuous change in conditions to be a-c phenomenon, and the steady-state condition to be d-c phenomenon, the reason for calling an *R-C* coupled amplifier an amplifier of a-c signals, becomes understandable. An *R-C* amplifier can transfer a change in a d-c voltage in exactly the same manner that it transfers an a-c voltage except that it is momentary while the change lasts. Once the changing state is replaced by a steady state, the a-c coupling system ceases to show any signs of the new condition.

The fact that triode tubes are used in the analysis of the *R-C* amplifier does not impose any limitation on the validity of the discussion. What was said applies equally well to all other types of tubes which see such service.

The Direct-Coupled (D-C) Amplifier

The direct-coupled amplifier differs in a number of ways from the *R-C* amplifier. The tubes may be the same, but the coupling between them and the distribution of the operating voltages differ greatly. Functionally, there is even a greater difference, which in the final analysis, is the most important factor. The d-c amplifier allows the transfer of a steady-state or a d-c voltage and the transfer of an a-c voltage with equal facility. This implies that the coupling capacitor *C*, which is a part of the link between the two tubes in the *R-C* amplifier and behaves as a blocking capacitor against the transfer of d-c voltages, is absent. Such is the case illustrated in Fig. 10-90. This is a representation of the fundamental Loftin-White form of direct-coupled amplifier.[3]

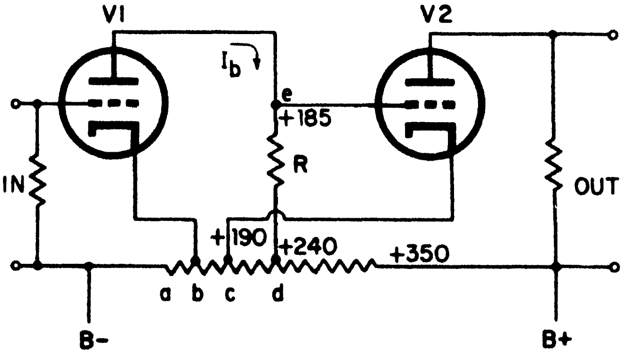


Fig. 10-90.—Schematic of simple d-c amplifiers.

Resistor *R* is the plate load for tube *V1* and at the same time the grid leak for tube *V2*. It would seem at first glance that the d-c voltage of *V1* is applied to the grid of *V2*. Such is the case but it is not effective in that way because of the manner of connecting the cathode of *V2* to the voltage supply system. If the cathode were connected to point *d* on the voltage divider, the control grid of *V2* would be negative relative to its cathode by an amount equal to the voltage drop across *R* which would numerically equal 240 − 185, or 55 volts. Because of the direction of current flow I_b, the control grid is negative relative to point *d*. Such a heavy bias would naturally make the tube inoperative. Therefore, the cathode of tube *V2* is connected to a point on the divider which makes that cathode *positive* relative to its control grid by an amount slightly less than the voltage drop across *R*. In other words, the voltage effective between *e* and *c* is the voltage drop across *c-d* minus the voltage drop *c-d*. According to the values shown, the voltage drop *e-d* amounts to −55 volts, whereas the voltage drop *c-d* amounts to +50 volts, thus making the control grid 5 volts negative relative to its cathode.

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[3] Loftin, E. H., and White, S. Young, "Direct-coupled detector and amplifier with automatic bias," *Proc. I.R.E.*, vol. 3, pp. 281-286, March, 1928.

In the absence of a signal at the control grid of tube *V1*, the plate voltage for that tube is equal to the voltage *a-d* minus the voltage *e-d*. A signal applied to the grid of *V1* will cause a change in the plate current of *V1*, therefore, a change in the drop *e-d*, and a change in the potential of the point *e* (the grid of *V2*) relative to its cathode. So, in the absence of a signal into *V1*, a state of equilibrium exists during which time the plate current of *V1* is constant; point *e* on *R* is at a fixed potential relative to its cathode, and the grid of *V2* is slightly negative relative to its cathode. This is illustrated in Fig. 10-91A and B at the left of the drawing where the no-signal conditions are indicated. Illustration (A) portrays conditions relative to the plate voltage of *V1*, and illustration (B) portrays the grid voltage of *V2*.

Let us imagine that a sine-wave signal of any frequency is applied to the grid of *V1*. We shall consider it to be of very low frequency since one of the reasons for the use of direct-coupled amplifiers is the convenience of obtaining excellent amplification of very-low-frequency signals. Only one cycle is shown because everything said in connection with this single cycle applies to any number of cycles.

The polarity of the signal on the grid of *V1* is such as to make the plate voltage of that tube rise. This can take place only if the plate current decreases and the drop across *R* is reduced, which of course changes the potential of point *e* on *R*, making it less negative as far as the grid of *V2* is concerned. Since the voltage drop *c-d* is substantially constant and the voltage drop across *R* has been decreased, the potential of the grid of *V2* becomes less negative than before and this change takes place in conformity with the instantaneous changes in the plate circuit of *V1*. During the first quarter-cycle of the signal voltage into *V1*, the plate voltage of that tube is rising and the grid of *V2* is becoming correspondingly less negative, following the changes in plate voltage exactly.

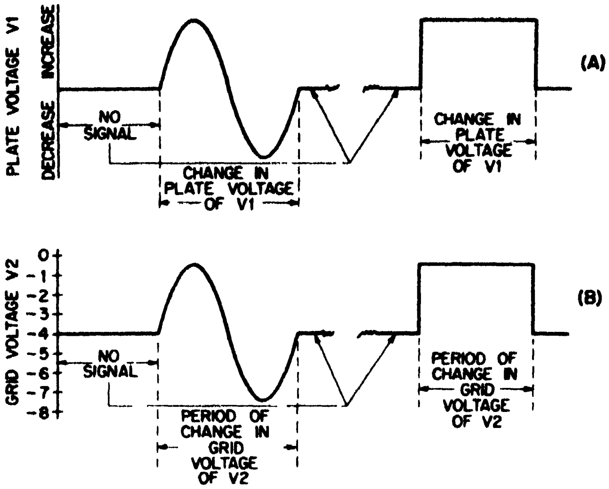


PLATE VOLTAGE V1
INCREASE
DECREASE
NO SIGNAL
CHANGE IN PLATE VOLTAGE OF V1
CHANGE IN PLATE VOLTAGE OF V1
(A)
GRID VOLTAGE V2
0
-1
-2
-3
-4
-5
-6
-7
-8
NO SIGNAL
PERIOD OF CHANGE IN GRID VOLTAGE OF V2
PERIOD OF CHANGE IN GRID VOLTAGE OF V2
(B)

Fig 10-91.—Illustrating the relation between the driving plate voltage and the driven grid voltage in d-c amplifiers.

When the plate voltage of *V1* reaches its maximum level, the grid of *V2* is at its maximum positive-going level, or, to put it differently, is least negative. As the plate voltage begins to fall during the second quarter-cycle of the signal (due to the gradually increasing plate current and increasing drop across *R*), the grid of *V2* follows this change by gradually becoming more negative. It reaches its original level at the time the plate voltage of *V1* falls to its original value in correspondence with the instant of zero signal on the grid of *V1*.

The reverse of that which has been described takes place during the next half-cycle; as the grid signal on *V1* becomes positive, the plate current of that tube increases, making the drop across *R* progressively greater and making the grid of *V2* progressively more negative relative to its cathode. In this fashion, the grid of *V2* follows the changes in plate voltage of *V1*, instantaneously assuming different levels of potential relative to its cathode, until the cycle is completed and the signal fed into *V1* is amplified and passed on to *V2*.

At first glance, the ability of the direct-coupled amplifier to amplify an a-c signal offers no special advantages, neither does it appear as anything significant. In reality it does offer advantages and we shall speak of them soon. For the moment, we desire to examine another capability of this system as an amplifier.

Continuing with Fig. 10-91 let us imagine that instead of an a-c signal input to tube *V1*, we apply a unidirectional voltage so that the plate current falls and the plate voltage of that tube rises. The input signal is a negative-going pulse and it is applied after a period of quiescence. The plate current falls suddenly to a new value and the plate voltage, therefore, rises in like fashion because the drop across *R* is reduced.

What happens in the grid circuit of *V2*? That grid follows the changes in potential at the point *e* on *R*; when the plate current fell, the drop across *R* decreased abruptly and the grid of *V2* became less negative just as rapidly, as shown in illustration (B) of Fig. 10-91. Then when the plate current of *V1* remained constant at the new value and the potential of point *e* (or the plate voltage of *V1*) remained constant at the new value, the grid of *V2* likewise remained constant at its new level of voltage relative to its cathode, as shown by the flat portion of the pulse in both illustrations of Fig. 10-91. This is a new state of equilibrium in the system for as long as the voltage conditions which we have described remain constant in the system. Of special interest is the fact that at the moment the effective potential between the grid and cathode of *V2* is much less than before, and to all intents and purposes this could just as readily as not be the new set of operating conditions for the system as a whole.

We know that is not the case, for after a while, the input signal to the grid of *V1* falls to its original value; the plate of that tube likewise returns to its original voltage because the increase in plate current through *R* increases the voltage drop across that element, and the grid of *V2* again becomes more negative, returning to its original status. Now, the performance of the amplifier when the input to *V1* was a pulse differs substantially from that of the a-c amplifier which we described previously. This can be seen by noting the action at the grid of *V2* in Fig. 10-89 and in Fig. 10-91 when a pulse was applied to *V1* in each case. When *R-C* coupling was used, energy was transferred from the plate of *V1* to the grid of *V2* only during the time the voltage was *changing;* the moment the change ceased, the grid of *V2* returned to its no-signal level. In the direct-coupled (d-c) amplifier, however, not only is the change transferred, but the steady-state condition following the change also is transferred. In other words, a d-c voltage has been amplified and transferred from one tube to the next. The new steady-state condition in Fig. 10-91B is the increase in plate voltage of *V1* and the less negative condition of the grid of *V2*. Since this is due to the d-c voltage applied to the grid of *V1* and it appears in amplified form at the grid of *V2*, a change in d-c level has been transferred from one tube to the other.

Now if we appreciate the full significance of the a-c signal which was amplified by the d-c amplifier of Fig. 10-90, it stands to reason that the amplification of a d-c signal is not limited to positive pulses. A negative-going pulse can be handled with equal ease, for all it means is that the drop across *R* (Fig. 10-90) is increased in conformity with the increase in plate current of *V1*, and the grid of *V2* becomes more negative as dictated by the characteristics of the signal in the plate circuit of tube *V1*.

So much for the basic principles underlying the operation of the d-c or direct-coupled amplifier, and the difference between it and the a-c or *R-C* coupled system. What we have said is by no means a complete exposition of the subject, there being much left unsaid because it is beyond the province of this book, and because explanations of different kinds of d-c amplifiers as used in practical instruments are given in Chapter 22. Sources of more elaborate information are given in the references contained in the bibliography. In the meantime several additional comments concerning d-c amplifiers are in order.

The first of these refers to the frequency range of such amplifiers. Their advantage lies in their ability to amplify d-c signals, therefore, extend down to zero frequency As to signals of very low frequency, 1 cycle or less, the d-c amplifier is the preferred system. The conventional a-c amplifier also is suitable for this type of signal, but the capacitor required for the coupling capacitance must, of necessity, be very large in size and this introduces a cost factor, but even more important than that is the increase in distributed capacitance and the resultant reduction in high-frequency response. Finally and perhaps even more important than the last named is the phase shift over the low-frequency range. The absence of the *R-C* combination in the d-c amplifier makes the system free from such effects.

Another advantage of the d-c amplifier is found in the ability to position the beam by the application of a positioning voltage in the signal channels. This means direct coupling between the deflection elements and the output amplifier as shown in Fig. 10-92, which is a greatly simplified version of the practical amplifier. It does, however, indicate one means for applying the positioning voltage. (Many more appear in Chapter 22.) This consists of creating an unbalance in the static level of plate voltages in one of the stages, in this example in the intermediate stage of amplification. This change in static level of one of the plate voltages is transmitted to the succeeding stages and appears in the output as a definite static unbalance which deflects the beam in one direction. The signal voltage is superimposed upon these voltages. The design of the system

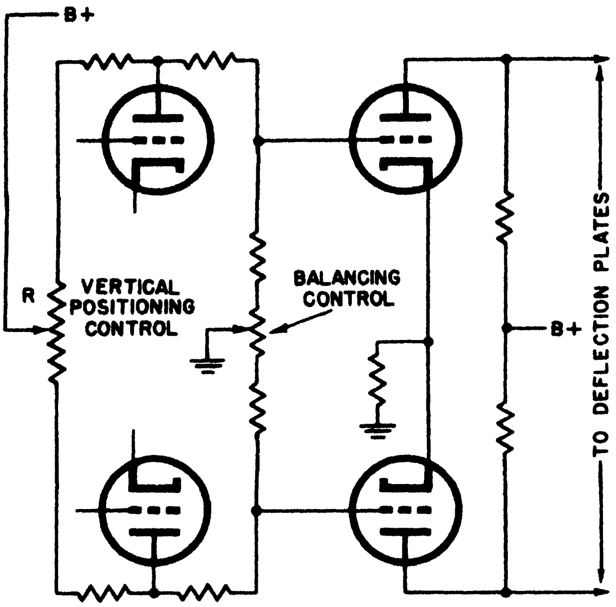


Fig. 10-92.—D-c amplifiers with positioning system (balanced) in the signal channel.

is such that the potentiometer *R* affords like plate voltages to both of the tubes which it controls when it is set at its midpoint. When set to either side, it applies a higher plate voltage to one or the other of the tubes, thus affording fixed displacement of the beam in either direction along the axis controlled by the amplifier. In this case, since we are speaking about vertical amplifiers, the positioning would be effective in the vertical direction.

The importance of d-c positioning via the amplifiers is that it affords the facility consistent with minimum effect on the frequency response. It must always be kept in mind that good frequency response is a prime requirement of all oscilloscopes. Even when the frequency range is limited, the location of variable controls in the amplifying channel is capable of adversely affecting the bandwidth of the amplifier. The ability to locate a positioning control as we have illustrated, which is only one arrangement of many, without displaying any effect on the bandwidth of the amplifier, is very important.

On the other side of the fence is the fact that the d-c amplifier is more critical in operation than its a-c counterpart. In fact, single-ended systems such as that shown in Fig. 10-90 are not suitable generally for purposes such as we are discussing, because the slightest amount of drift in the operating potentials is transmitted through the system just as if it were a signal. To overcome this, recourse is made to push-pull systems for all stages. A satisfactory balance is thus attained, although, even under such circumstances, it may be necessary to utilize a regulated power supply for the stages ahead of the output stage. The need for accurate balancing is met by the inclusion of a number of balancing controls which arrange the correct distribution of operating potentials. However, it does introduce some maintenance problems. Replacement of tubes and resistive elements is, of course, possible when one or more of these components fail, but they cannot be replaced with complete abandon. While it is true that the balancing controls provide a latitude of tolerance in the values of the resistors and in the characteristics of the replacement tubes, there does remain a limit which must be observed. Therefore, when replacement of a tube is necessary in the balanced stage of a d-c amplifier, it is highly recommended that both tubes in the stage be replaced; moreover, if they are individual tubes, they should be of like characteristics. Of course, when the stage employs a dual-section tube, replacement is simple, but even then, the characteristics of the tube selected as the replacement should match the characteristics of the original tube most closely. Invariably replacements of this kind demand readjustment of the balancing controls, and if the original degree of stability is not attainable, it means that the values of the replacement units exceed the tolerances which can be corrected by the existing balancing controls.

Practical D-C Amplifiers

When an amplifier is rated as a direct-coupled system, it means that the coupling capacitance has been eliminated in *all* stages, inclusive of the coupling to the deflection plates. A combination of a-c coupling in one stage and d-c coupling in the remaining stages does not constitute a direct-coupled system. If anything, it is an a-c system even if the *R-C* coupled stage is singular in comparison with several direct-coupled stages.

Occasionally the vertical amplifier in an oscilloscope is stated as affording a choice of either a-c or d-c amplification. This invariably means a choice at the input point and seldom if ever in the stages of the amplifier. What we mean is shown in Fig. 10-93. The amplifier proper (not shown) is a d-c coupled system, but the input circuit feeding the attenuator is switch-controlled for either a-c or d-c signals. In one position, the grid system of the input stage is connected directly to the "high" terminal of the input system, since the switch position shorts the coupling capacitor *C*. This is the position marked *DC*. When the switch is set to the *AC* position, the coupling capacitor is in the circuit. The

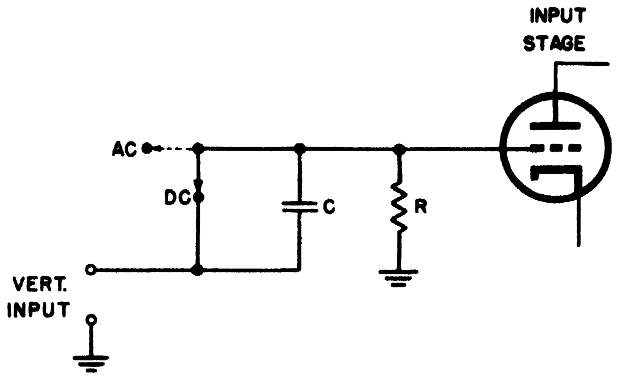


Fig. 10-93.—Input selector switch of a d-c amplifier.

system we show is only a part of the complete circuit of these units, but it is sufficient to illustrate the point.

There is no such thing as an a-c d-c amplifier. If it is an a-c amplifier, it is not suitable for the transfer of d-c signals, whereas if it is a d-c amplifier, it automatically includes a-c voltages within its frequency rating.

D-C Coupling and Square-Wave Amplification

It is possible that some confusion may arise in the mind of the reader concerning d-c coupling and square-wave amplification. The latter is possible with direct and *R-C* coupled systems, but the disadvantage of the latter when low fundamental frequencies are involved is, as was mentioned earlier, the need for high values of coupling capacitance, and also, the matter of phase shift. When the square-wave fundamental frequency is several cycles or less, d-c amplifiers are much to be preferred.

Another point of interest relates to the possible presence of a d-c level as the base of the square wave or pulse. As we have shown, only the d-c amplifier is capable of properly passing such a signal, therefore, direct coupling in the vertical amplifier of the oscilloscope is a must if it is to indicate the d-c level as well as the shape of the signal.

Phase Inverters and Push-Pull Systems

The use of single-ended or single tube stages in most of the amplifier diagrams that we have discussed is not intended to infer that the vertical amplifiers in oscilloscopes used such arrangements. Their use is common, to say the least, but the push-pull arrangements are more general, especially in the more elaborate instruments. In these units, they are to be found in all the stages, whereas in the instruments which make use of the single-ended stages, only the final output stage utilizes the push-pull circuit, and then not in every instance. In those cases where the amplifier is a combination of single-ended stages and a push-pull output stage, phase inversion is necessary in order to feed the output stage with two signals which are 180° out of phase.

The operation of the push-pull stage and the phase-inversion stage is the same as that for the more commonplace audio amplifiers. If there is anything of special significance, it is found in the constants selected so as to minimize the deterioration of the band-pass capabilities of the amplifier in the phase inversion stage. In the usual audio amplifier with a pass-band of perhaps 10 kc, neither phase shift nor the loss of frequencies above 10 kc is of importance; but in the wide-band oscilloscope amplifier, both are pertinent to good performance. The possible arrangements for attaining phase inversion are numerous, entirely too many to allow illustrating all of them. Those which are representative of generalized art are shown; these are the basis of the numerous arrangements to be found in oscilloscopes.

If we seek the phenomenon of phase inversion, it is found in the operation of the usual three or more element tube, namely in the phase relations between the control grid, plate, and cathode circuits. The signal voltage at the plate of a vacuum tube is 180° out of phase with the signal voltage at the grid; the signal voltage at the cathode is in phase with the signal voltage at the control grid, and 180° out of phase with the signal voltage at the plate. With these three cardinal conditions as a basis, especially the last named, it is not difficult to visualize two signal voltages 180° out of phase being secured from a single tube. An amplifier of this type is sometimes referred to as a "paraphase" amplifier, the coined word being the equivalent of "a pair of phases" as applied to two signals. One of these signals is secured from the plate system, and the other from the cathode system. This is illustrated in

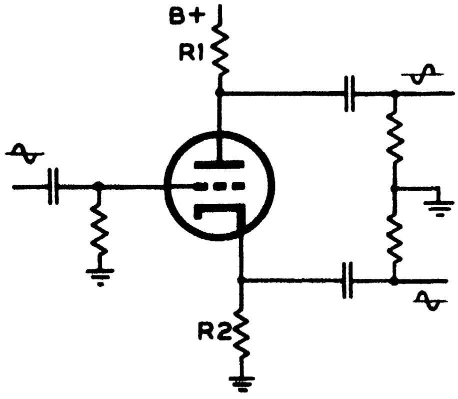


Fig. 10-94.—Schematic of a simple phase inverter.

Fig. 10-94. The values of *R1* and *R2* are the same, and since the same signal current flows through both, equal voltage drops will develop across each. These will differ in phase by 180° because an increase in plate current will make the plate less positive than it was before, whereas it makes the cathode more positive than before the increase. Gain in an amplifier of this type is of secondary importance, since the primary requirement is two output signals, of like magnitude but differing in phase by 180°.

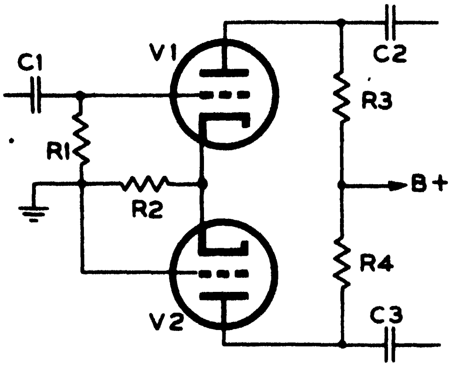
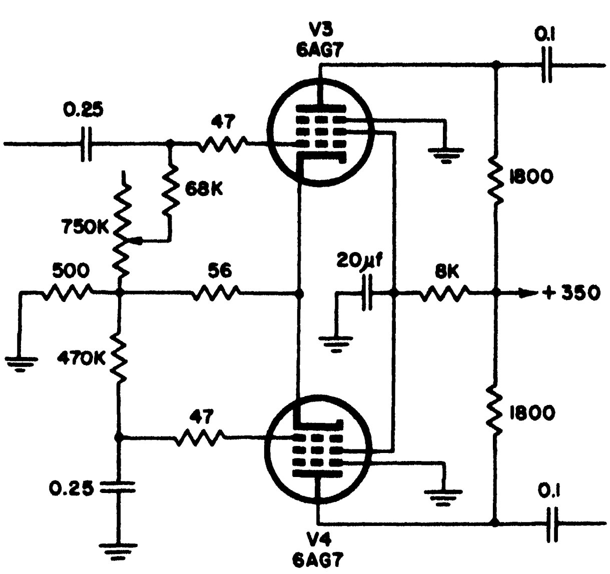


Fig. 10-95.—Schematic of a cathode-coupled phase inverter.

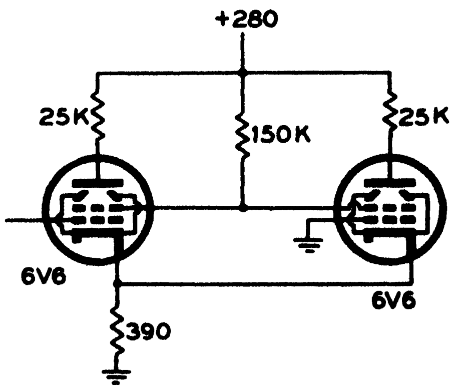
One version of cathode-coupled phase inversion in the output stage of an oscilloscope amplifier in shown in Fig. 10-95. All of the frequency-compensating networks have been omitted in order to show the operation of the system more clearly. The signal is fed into the push-pull stage from a single-ended stage via the coupling capacitor *C1*. With bypassing of cathode resistor *R2* omitted, a signal voltage is built up across this element and ground, and it has exactly the same polarity and shape as the signal fed into the grid of tube *V1*. With a common connection between the cathodes of tubes *V1* and *V2*, and with the control grid of *V2* grounded, any rise in the potential of control grid *V1* will cause a rise in the potential of the cathode of *V1* and a corresponding rise at the cathode of *V2*, and, since the control grid of the latter is connected to the low side (ground) of the cathode resistor, it is made correspondingly negative relative to its cathode. Thus, when one grid is negative, the other grid goes positive, so that the basic requirements for push-pull action have been obtained without resorting to the use of a separate tube for the phase inversion process.

While it is true that a condition of unbalance does exist in the grid circuit, which is essential to its operation, the proper choice of the constants for the different elements results in a substantially uniform output from each stage with the two signals 180° out of phase with each other. A typical oscilloscope amplifying stage of this kind is symbolized in Fig. 10-96. The frequency-compensating networks have been omitted from the plate circuit. The essential difference between this input and the one previously described is that the connection to the control grid of tube *V4* is above ground, but its junction is along the low side of the common cathode resistor so that it is made negative relative to its cathode, by a rise in grid voltage at *V3*.



Courtesy RCA

Fig. 10-96.—Typical oscilloscope phase inverter.

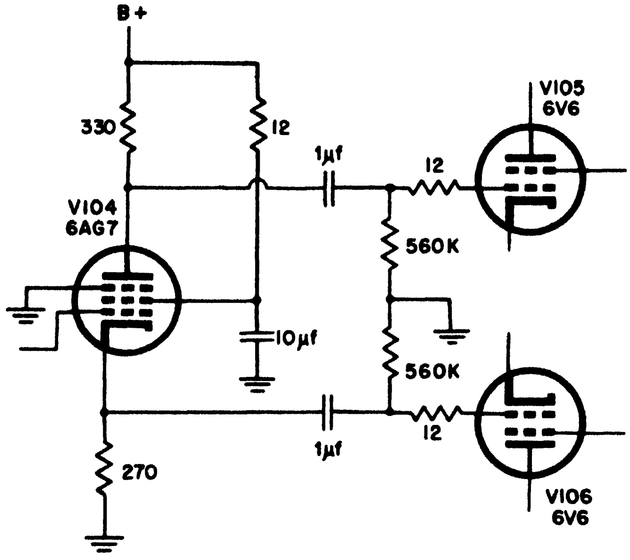


Courtesy Du Mont

Fig. 10-97.—Typical grounded-grid oscilloscope amplifier.

A specific example of the grounded-grid system is shown in Fig. 10-97. This is found in the Du Mont 208-B and in other units. As in the previous case, the frequency-compensating networks have been omitted.

A commercial example of phase inversion which corresponds to the basic system of Fig. 10-94 is shown in Fig. 10-98A. Tube *V104* is the phase inverter, and tubes *V105* and *V106* are the push-pull stages for which the two 180°-out-of-phase signals are required so that push-pull action can take place in the plate circuits.



Courtesy RCA

Fig. 10-98A.—Commercial example of a phase inverter corresponding to the basic type of Fig. 10-94.

As can be seen in the schematic, *V105* receives its signal voltage from the plate of *V104* via the coupling capacitor. This affords a signal of a certain instantaneous phase ; for the moment, we can assume it to be positive. At the same instant, the control grid of *V106* receives its signal from the top of the phase-inverter cathode resistor. As we stated before, signals at the cathode and plate of a tube are 180° out of phase, so that the signal fed at the same instant to the control grid of *V106* is negative. Thus the push-pull grids receive signal voltages which are 180° out of phase.

Now, in addition to the phase requirement is the need for like values of signal voltage at the two push-pull grids. The attainment of a signal voltage across the cathode resistor requires that it not be bypassed to ground, which is the case for the plate-load resistor. Now as the consequence of the absence of cathode bypassing, the development of a signal voltage across that resistor is in effect equivalent to a corresponding reduction in the signal voltage at the control grid of that tube. This signal is effectively out of phase with input signal voltage and thus tends to reduce the signal voltage effective at the grid. This corresponds to degeneration which very materially reduces the output signal from the tube as a whole. The actual signal effective at the control grid of tube *V104* is the difference between the input signal and the signal appearing across the cathode resistor. This means that the signal across the cathode resistor (which is one part of the output voltage) never can be greater than the original input signal, and the tube as a whole is not used as a means of gaining amplification.

In the usual circuit of this type, the plate-load and cathode resistors are of equal value. Since the same plate current flows through each, the voltage across each will be the same. However, because of the wide pass-band of this amplifier, unbalance would exist at the high frequencies because of unequal shunt capacitance distribution. At the low frequencies, unbalance would exist because of incomplete screen bypassing. In order to provide equal output voltages over a relatively wide range of frequencies, the plate and cathode resistors in the circuit of Fig. 10-98A are made unequal to compensate for the unbalancing effects.

Numerous other versions of phase inverters are to be found in modern oscilloscopes, but most of them are based on the two systems shown in Figs. 10-94 and 10-95. There are, of course, many other possibilities which may be used, and if they are not to be found in the oscilloscopes shown herein, they may possibly appear later.

Advantages of Push-Pull Amplification

The reasons for using push-pull amplification are quite numerous. Perhaps the most important of these is the desire to establish a *balanced* condition between the deflection plates and the accelerating anode. Whereas in the unbalanced (single-ended) arrangement, one deflection plate varies in potential relative to the accelerating anode as the deflection voltage is applied, in the balanced system (push-pull) one plate is made negative just as much as the other plate is made positive. This results in a uniform field between the deflection plates and surrounding elements which are at different potentials from it, and the beam remains in focus as it completes its excursion across the screen surface. Single-ended systems, on the other hand, are prone to display defocusing as the extreme limits of the beam excursion are reached.

Another advantage gained by this symmetrical method of feeding the deflection voltages is the attainment of a greater degree of linearity in the deflection. Since both plates swing the same amount in the unlike potentials, the forces acting on the electron beam are the same in both directions and the trace is more linear relative to its undeflected position. In connection with linearity of the trace, the push-pull system tends to minimize even harmonic distortion and its contribution to nonlinearity. The presence of odd harmonic distortion may influence the characteristics of the trace, but it will not impair the linearity of the deflection.

In addition, the use of a push-pull output stage affords twice the signal voltage swing of one tube for the same supply voltage. This comes about as the consequence of the action that as one plate swings positive, the other plate swings negative by an equal amount, and the total swing is the sum of these two. This is important when large dimensioned screens are used, or, for that matter, with any size screen, since it affords a much greater range of output voltages for the minimum supply voltage. It is a distinct advantage when a certain portion of the trace must be investigated and greatest readability is attained by spreading the trace in either or both directions.

Phase Relationship Between Input and Output Signals

In describing the association between the polarity of the input signal voltage and the direction of deflection, we must assume a standard condition, namely that a *positive* swing in input signal voltage will cause an *upward* direction of deflection, and a *negative* swing will cause a *downward* deflection. Since the phase relationships in a conventional amplifier stage are such that the grid voltage will be inverted in phase in the plate circuit, the number of tubes which are in the path of the signal generally determine the phase of the output relative to the input signal. The importance on this relationship is found in the direction of deflection on the screen of the cathode-ray tube.

In order that the signal voltage at the output of the vertical amplifier may produce upward deflection for a positive input signal, proper connections must be made to the vertical deflection plates, relative to the orientation of the cathode-ray tube in the oscilloscope. Fig. 10-98B(1) shows a positive-pulse input signal to a simplified single-stage amplifier. The output of the amplifier is connected to the vertical deflection plates of the cathode-ray tube, which is positioned so that a *positive*-going signal at *V1* will deflect the beam upward. Since the amplifier inverts the pulse, the *negative*-going plate signal is applied to deflection plate *V2*. If a two-stage amplifier is used, as in Fig. 10-98B(2), the output will be in phase with the input signal. Therefore, to produce upward deflection for a positive-pulse input, the output of the amplifier is connected to deflection plate *V1*.

Advancing to the generalization of the number of stages and the phase conditions, let it be said that amplifiers with an even number of stages will invert the phase of the signal between the input and the output circuits, whereas amplifiers which employ an odd number of stages will generally repeat the phase of the input signal at the output circuit. The use of positive pulses for the input signal is not a limitation of the statement; it is true regardless of the instantaneous polarity of the input signal voltage. Moreover, it is equally true in either a-c or d-c amplifiers.

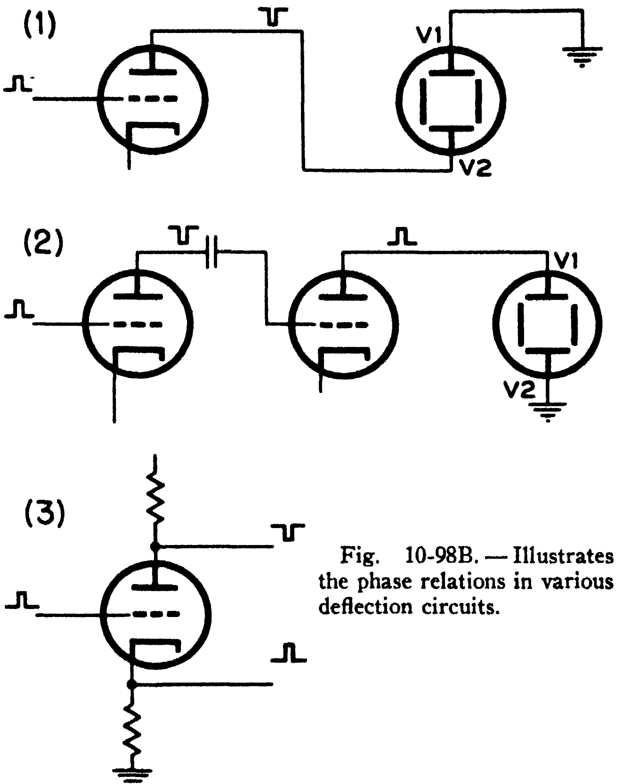


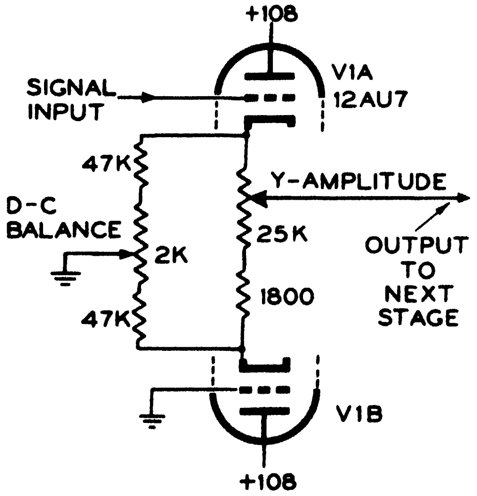
Fig. 10-98B. — Illustrates the phase relations in various deflection circuits.

It is possible, however, to have a number of stages in an amplifier, and yet not obtain the phase inversion described. This is shown in Fig. 10-98B(3), wherein by using phase inversion, a single tube accomplishes the phase functions of two individual tubes; the signal voltage is taken off at the cathode resistor. At this point, the grid and cathode voltages are in phase, whereas the signal voltage at the plate is 180° out of phase with the signal voltage at the cathode.

Because of the possibility of such phase inverter systems in an amplifier, snap judgment of the phase of the signal at the input to the amplifier cannot be based on the direction of deflection as seen on the screen, unless the manufacturer of the oscilloscope indicates the relationship between the polarity of the input signal and the screen deflection. Not all oscilloscopes deflect upward with positive input signals, although the majority do.

Cathode-Follower Input System Circuits

The earlier reference in this chapter to the cathode-follower input system concerned the simpler versions; at this time, we desire to show a more elaborate system which is used in some of the special-purpose oscilloscopes. The basis of operation conforms with the previously described cathode-coupled phase inversion process, except that in this instance the two sections of the input tube act together to furnish a signal to the succeeding stage in a d-c coupled system. This circuit, shown in Fig. 10-99, is a simplified version of the cathode-follower input in the Du Mont 304 oscilloscope and other related models.



After Du Mont

Fig. 10-99.—Simplified schematic of a cathode-follower input system.

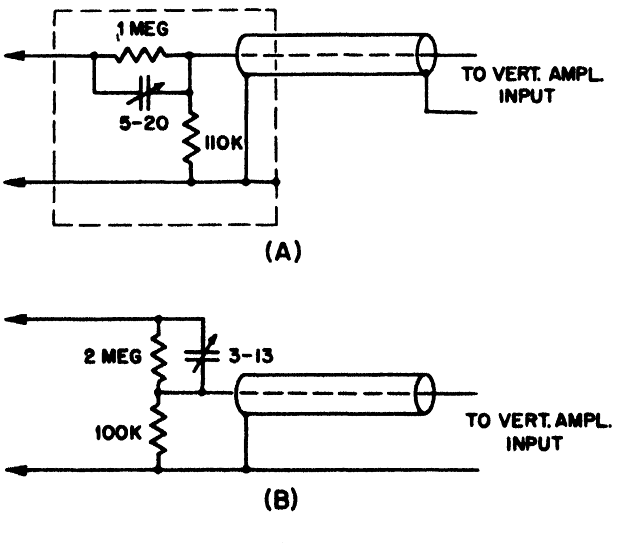
The d-c balancing potentiometer is used to adjust the d-c levels so that they will be alike at each end of the *Y*-amplitude control. The output to the succeeding stage is unbalanced and direct-coupled, but the d-c level at the arm of the *Y*-amplitude control remains constant. Triode *V1B* serves only to keep the d-c output balanced.

Input Probes and Attenuating Cables

Since the interpretation of a basic oscilloscope must of necessity be flexible, it is necessary to associate with the vertical amplifier certain accessories, which, while not found as a part of every oscilloscope within the basic category, may be found with some which can rightfully be placed in that classification. Such accessories are ***input probes*** and ***attenuating cables***. In the main, these are one and the same thing, although we must recognize that in one or two isolated cases the input probe may be a different sort of device than the one we will describe. This special version will be discussed later. For the moment, let us consider the general version of the probe.

Whether we call the device a probe or an attenuator cable, it has two specific manners of behavior; one is its desired purpose, the other is undesirable and unavoidable, although put to use. To understand these uses, let us refresh your memory concerning the input impedance of the usual vertical amplifier. It presents both resistance and capacitance to whatever circuit may be connected across it. The ideal condition would be the highest resistance and the lowest capacitance. Although these are made as high as possible and as low as possible, respectively, the capacitance may not be as low as may be desired. The input capacitance of the average vertical amplifier may range from about 30$\mu\mu$f to 60$\mu\mu$f, and the resistance which it presents may be from as low as 500,000 ohms to as high as several megohms.

Sometimes these constants are such as to present a problem when the oscilloscope is used with certain equipments; the input resistance may be high enough, but the input capacitance may be so high relative to the device which is to be investigated as to severely load the latter, or to modify the character of the voltage being observed. To avoid such conditions, some manufacturers make available special cables; these may be called probes because of the physical shape of the unit at the end of the cable, or simply low-capacitance attenuator cables. These are intended to be used as the connecting link between the device being checked and the regular input to the vertical amplifier.



(A) Courtesy RCA, (B) Courtesy Browning

Fig. 10-100.—Schematics of two typical input cables.

The constants of two of these probes or cables which are typical of commercial devices appear schematically in Fig. 10-100. The advantage which it affords is a major reduction in input capacitance of the vertical amplifying system and sometimes a major increase in input impedance to as high as 5 megohms. Such a probe used with a system rated at 50$\mu\mu$f input capacitance reduces the capacitance of the input circuit to from 10 to 15$\mu\mu$f. This modification is *not* effective on the actual *input* of the *amplifier*, but *exists* at the *input* of the *probe;* in effect, the actual input circuit of the vertical amplifier is isolated from the device which is connected to the input of the probe.

Such an improvement is not accomplished without some sacrifice, this being the attenuating action of the probe, hence the alternate name of attenuator cable. As a rule, such cables attenuate the signal by a 10:1 or 20:1 ratio, so that the use of the cable in order to reduce the capacitive loading is limited to those instances when adequate signal level exists. While it might be advantageous to use such a probe or cable in every case, it is frequently impractical because insufficient gain exists in the vertical amplifiers to offset the loss in the cable. On the other hand, many applications demand such a cable, and it was designed for that purpose.

Another version of an input cable supplied with some oscilloscopes is known as a *direct* cable, which refers to the passage of the signal from the point where the cable is connected to the input of the amplifier, rather than to the use of d-c coupling. Such a cable affords the facility of a shielded extension of the input of the oscilloscope to the point desired, but with a substantial increase in the effective input capacitance of the system. The added capacitance is that presented by the cable, and may amount to from 50 to 200$\mu\mu$f.

Still another version of the oscilloscope probe, and definitely different from that which we have illustrated, is the cathode-follower unit. This probe contains a miniature tube connected as a cathode follower. The tube is located at the tip of the probe, the probe tip being connected to the control grid of the tube and the ground lead being connected to the grounded end of the cathode resistor, which is located in the probe. The input to the vertical amplifier is taken off at the top end of the cathode-follower cathode resistor and this feeds into the attenuator system inside of the oscilloscope unit. Such a unit affords a substantial input resistance and a very low input capacitance. The power for the heater of the cathode follower also is contained in the interconnecting cable between the probe and the chassis of the oscilloscope. The circuit is like the cathode follower described earlier as a part of the attenuator system.

The oscilloscope end of these special cables and probes terminates in a special type of connector, such as the Amphenol or Cannon variety, and this is intended to be plugged into a corresponding female socket accessible on the panel. Sometimes the binding-post connections which serve the input of the vertical amplifier also join with the "probe input" connector in the oscilloscope, so that either may be used with minimum changes. It must be remembered, however, that the binding-post junction exists in the event that the "high" terminal is connected to some device, because the lead would short out the attenuating system if the "high" side of the attenuator probe also made contact with the same signal point.

Some oscilloscopes provide a switch at the input to the vertical amplifier which selects between binding-post input and probe input. In the latter case, the "high" binding post for the vertical amplifier is disconnected from the input of the vertical amplifier.

CIRCUIT FEATURES OF THE HORIZONTAL AMPLIFIER

It is possible to treat the horizontal amplifier as completely as we have its vertical counterpart, but if this were done, many subjects common to both would be repeated. It therefore seems most expeditious to simply refer back to the vertical amplifier for those details which are common to both. It is only natural that many circuit conditions and features found in the X-axis amplifier should be the same as those in the Y-axis amplifier.

Horizontal Amplifier Bandwidth

There was a time, during the early days of oscilloscope use, when certain definite distinctions were made between the vertical and the horizontal deflection amplifiers. They differed in gain as well as frequency response in almost every instance, because of the comparatively limited use of the device as a whole. Although the above is still generally true, the horizontal amplifier in many instances possesses performance characteristics which very closely approximate those of the vertical amplifier. If a line is to be drawn which *may* indicate where these two amplifying systems depart in pass-band ratings, we might say that it begins with the oscilloscopes which have vertical amplifiers with band-pass characteristics of about 1.0 Mc and higher. In elaborate devices of this type, the horizontal amplifiers do not equal the vertical amplifiers in performance specifications. In those oscilloscopes with top frequency ratings below 1 Mc, both the horizontal and vertical amplifiers frequently offer similar band-pass response.

The factor of economy is a very important one in all of these units, especially when the probable uses are taken into account. As we described earlier, the general run of waveforms which are generated by the time-base oscillator make less demands upon the related amplifying system for proper reproduction than signals which may be fed into the vertical system. Therefore, the added cost involved in making the horizontal amplifier design equal to the vertical amplifier performance above 1 Mc is most certainly not worthwhile. In those devices which might have more general appeal for limited use, such as maintenance applications, there is no need for elaborate designs beyond a certain point; this is reached when both the horizontal and vertical amplifiers have band-pass characteristics which will accept signals up to from perhaps 100 kc to about 500 kc, depending on the specific design of the instrument.

Horizontal Amplifier Input Impedance

As far as the input impedance is concerned, the problem is much simpler and no special effort is required in order to make the input-circuit constants of the horizontal amplifier equal to those of the vertical amplifier, even if the remaining performance ratings of the former are inferior to the latter. Consequently, while the demands of the time-base oscillator are less important relative to the input impedance of the horizontal· amplifier, it will be found that this operating constant very closely approximates, if not equals, the input constants of the vertical amplifier in virtually all oscilloscopes.

X- versus Y-Amplifier Circuits

There are numerous points of similarity between the X- and Y-axis amplifiers. These do not warrant much discussion, although we might mention in general terms what they are. For example, both single-ended and push-pull stages are to be found in the horizontal amplifier and their basic design is the same as that of the vertical system. The degree of frequency compensation may not be as great, but since this is not required for the proper transfer of the time-base voltages to the deflection plates, it is carried out only to a limited extent. As to the reasons for the use of any one system, whatever may have been said about the vertical amplifier also applies to the horizontal amplifier, because, in the final analysis, the ultimate destination of the output of the horizontal amplifier is beam deflection. The need for the retention of proper focus, freedom from effects of supply-voltage variations, maximum beam swing for minimum supply voltage, in fact, all of the operating conditions desirable in the vertical system are duplicated in the horizontal-deflection system by circuits which are virtually the same in both.

The organization of these amplifying systems does not differ greatly from that of the vertical amplifiers. The input systems may or may not use step attenuators; this again is a matter of individual oscilloscope design. Usually the arrangement found in the Y-amplifier is incorporated in the X-amplifier, simple or elaborate, whichever it may be.

Concerning the performance of the horizontal amplifier, one item warrants attention, frequency discrimination and change in waveshape due to the attenuator setting. As a system, neither the horizontal nor the vertical amplifier can distinguish between signals of different waveshape which are fed into them. Therefore, any system which can affect the composition of a nonsinusoidal signal can do so regardless of where it may be used. Thus, the attenuator used in the horizontal amplifier merits consideration in connection with its effects on signals fed into it.

If we imagine the horizontal amplifier being used to feed a square wave or similar pulse to the horizontal deflection plates, everything we said relative to the requirements of the vertical amplifier is true. In fact, the probability exists that because of the limited response of the horizontal system, the range of fundamental frequencies which may be passed into the horizontal amplifier and still be free of distortion will generally be less than that for the Y-amplifier. Since this is more apt to be true as the bandwidth of the vertical amplifier increases, it is most important to recognize the limitation of the horizontal amplifier. Otherwise, degraded quality of reproduction may be experienced.

The fact that the *primary* function of the horizontal amplifier is the amplification of sawtooth waves greatly determines the response of the amplifier. You will recall an earlier remark that acceptable square-wave reproduction requires the presence of approximately ten *odd* harmonics, whereas the sawtooth wave requires a total of ten harmonics of the fundamental for an acceptable facsimile. This explains why the frequency range of the sawtooth oscillator in an oscilloscope is so high in comparison with the square-wave frequency response of the amplifier.

It is interesting to note that probe and attenuator cables are not used with horizontal amplifiers; they are intended primarily for the vertical system.

The Horizontal Gain Control and Sweep Amplitude

Although we have not yet discussed the time-base oscillator other than in a perfunctory manner earlier in this chapter, we can at this time examine the behavior of the horizontal gain control on the sawtooth wave fed to the horizontal-deflection plates. The effect we have in mind is the change in shape of this voltage during its passage through the horizontal amplifier, and not any effect which the horizontal-amplifier gain control may have on the time-base oscillator.

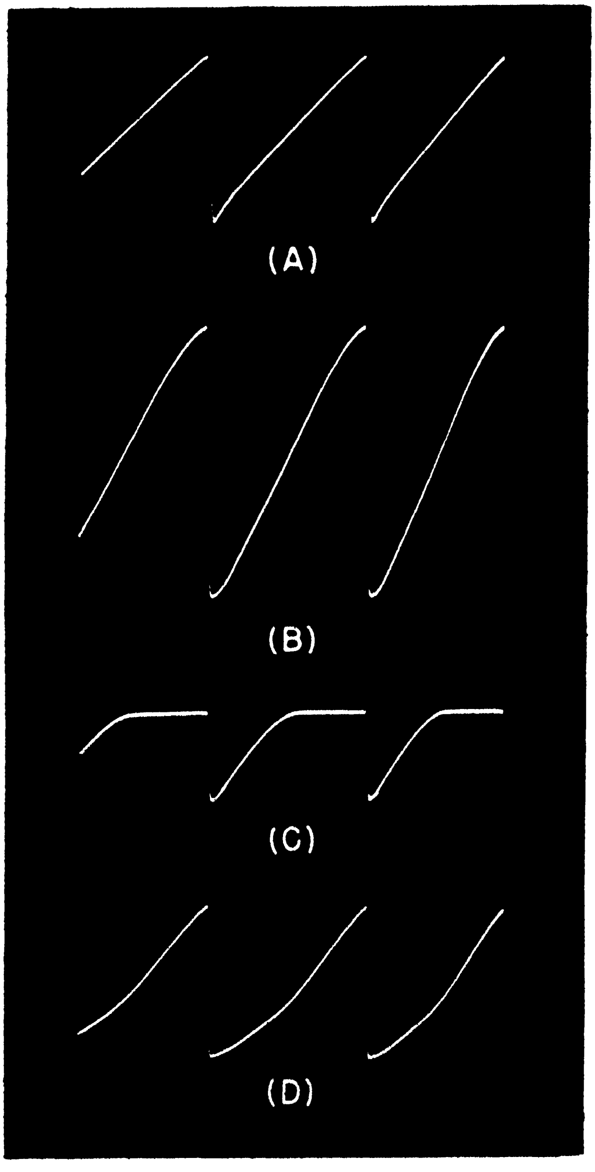


Fig. 10-101. — (A) and (B) substantially linear sawtooths resulting from ordinary positioning control settings. (C) and (D) non-linear sawtooths resulting from extraordinary positioning control settings.

The reason for this discussion is the equipment specification which states the possible expansion of the sweep amplitude relative to the dimensions of the screen. Conceivably this might be considered a part of the time-base oscillator discussion, but since the output of this oscillator is constant, and the sweep amplitude is a function of the amplifier gain control, we treat it here. The expansion of the sweep amplitude beyond the limits of the screen is an important performance capability of every oscilloscope, because it is the means of studying one portion of the trace, which otherwise might be too small. Some oscilloscopes afford a sweep amplitude as much as six times the screen diameter, whereas other oscilloscopes may limit this expansion to twice the screen diameter.

The ability to expand the sweep beyond the limits of the screen diameter and still retain the desired characteristics of the sawtooth is a function of the horizontal amplifier design; the requirement is that the input stage be capable of accepting the *full* swing in signal voltage which corresponds to the expanded sweep on the screen.

Oscilloscope design is such that the expansion of the sweep within the rated limits as set forth in the specification, as for example four or five times the screen diameter, can usually be attained without difficulty or alteration of the linearity of the trace. As a matter of fact, as shown in Fig. 10-101, which is typical of most instruments, the portion of the sweep which remains on the screen regardless of the settings of the position control usually will be sufficiently linear to develop a satisfactory trace for study purposes. This is shown as (A) and (B) in Fig. 10-101. It is only when extreme conditions are created, that is, when the horizontal-positioning voltage (also applied to the horizontal amplifier) tends to move the spot completely off the screen or to its extreme limits, that the trace is badly distorted and made nonlinear. This is shown as (C) and (D) in Fig. 10-101.

The important thing to bear in mind in connection with these facts is that each oscilloscope is designed to afford a certain range of sweep amplitude via the horizontal gain control, and that this information is a part of the specifications. If these limits are exceeded, the horizontal amplifiers will be overloaded, but if the sweep amplitude is not made too great, there will remain a portion of the scanning trace which will be linear and will be suitable for the development of a pattern. This will be true only if the positioning control is set so as to leave a trace across the screen. If the positioning control is advanced so far that only a segment of the trace is on the screen, the trace will be badly distorted. The ends of the trace at either the right or left edges of the screen are the overloaded portions of the sweep voltage, and will not be suitable as time-base voltages.

It might be advisable to mention that much better results will be obtained at all times if the midportion of the sawtooth time-base voltage, rather than the limits of the voltage swing, is used for observation of the trace on the screen. This is why the recommendation is made to always set (if possible) three cycles of a voltage on the screen and to use the center cycle for study purposes. Such a trace may be expanded to the maximum and if only the middle cycle remains on the screen, the other two being beyond the limits of the screen, adequate linearity of the sweep will prevail and critical study will be possible.

All of this leads to the conclusion that the action of even the simple horizontal gain control on the sawtooth time-base wave is not as critical as in the case of the square-wave voltage applied to the input of the vertical amplifier. But, if the horizontal amplifier is used for the amplification of a square wave or a rectangular pulse, then all that was said about the frequency-discriminating action of the vertical-amplifier gain control is applicable to the horizontal gain control. In making these statements we are not denying the possible action of frequency discrimination in the horizontal amplifier. Earlier we stated that frequency compensation is found in both amplifiers and it is always important to maintain the proper characteristics of the sawtooth sweep voltage. The fact that its frequency demands are not as severe as for other forms of nonsinusoidal waves does not lessen the need for the presence of the range of frequencies required for suitable wave reproduction.

So much for the general details about the horizontal amplifier. Other significant features of this section of the oscilloscope will be dealt with separately after the discussion of the time-base signal. The correlation between these systems can best be accomplished after the sweep system has been analyzed.

THE TIME-BASE GENERATOR (REPETITIVE SWEEP)

Since Chapters 9 and 11 cover the circuits and operation of the variable time-base oscillator, it is unnecessary for us to dwell at great length upon either the fundamental action or the specific circuits which are used. We shall focus our interest upon the general conditions of performance in the commercial oscilloscopes.

The variable-frequency sweep circuits found in oscilloscopes may be classified in two groups. One of these is the ***recurrent or repetitive*** sweep, whereas the second category is the ***nonrepetitive or triggered*** sweep. We have shown that in the basic oscilloscope the variable-frequency time-base oscillator produces the first type of sweep voltage. This is the sawtooth voltage which repeats its cycle of operation as long as the operating voltages are applied to the circuit, and in so doing will repeat the presentation of the signal trace as long as the voltage to be examined is fed into the vertical amplifier.

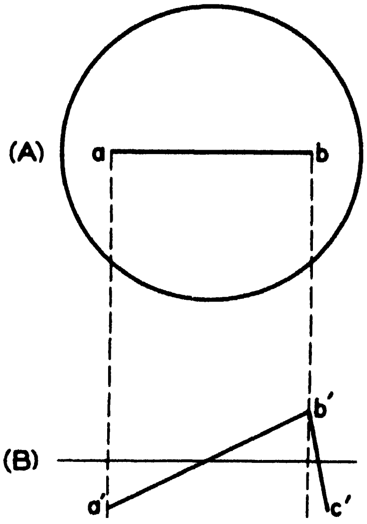


Fig. 10-102.—(A) is the trace of a sawtooth deflection voltage, and (B) is the sawtooth voltage plotted against time.

By itself, the time-base oscillator voltage appears as a straight line across the face of the tube, oriented horizontally, as shown in Fig. 10-102A. If the shape of this voltage were examined, it would appear as in Fig. 10-102B. The correspondence between the trace on the screen due to the application of only the sweep voltage, and the portion of the sweep voltage cycle which is productive of the time-base trace is shown by the dotted lines which join curves (A) and (B). From this, it becomes evident that the movement of the spot across the screen, in this case from left to right, is a function of the electrical circuits between the output of the horizontal amplifier and the two deflection plates. The start of the sweep point *a* corresponds to the most negative level of the sweep voltage *a′*, and the end of the sweep *b* corresponds to the most positive level *b′* of the sweep voltage. Whether the sweep will move from the left to the right, or in the reverse direction, is dependent upon which of the two horizontal plates is positive relative to ground. In virtually all oscilloscopes, the polarity of the deflection plates is such that the time-base voltage trace begins its excursion at the left side of the tube face and advances toward the right.

Having traveled across the screen, the spot must return to its starting point so as to begin a new sweep. This is the retrace and takes place during the time interval *b'-c'* in Fig. 10-102B. This is not evident on the screen when vertical deflection is absent because it follows the same path as the forward trace. It is, however, very much in evidence in most instances when vertical deflection is applied and its path is determined by the portion of the cycle of the vertical deflection voltage which prevails during the retrace period. Referring to (B) in Fig. 10-102 the *cycle* of the sweep voltage is the time between *a'* and *c'*, or the forward trace plus the return trace. The interest lies in the relative time intervals of *a'-b'* and *b'-c'*, or the ratio of the time of the forward trace and the time of the retrace. As a general rule, the retrace time consumes from about two per cent to ten per cent of the total cycle time, sometimes more.

Obviously, the less the retrace time relative to the forward trace time, the less the portion of the display which is lost for observation or study. Since the retrace time is always finite, and since this period is not the same in all cases, sometimes not even being known because it is not stated in the literature which describes an oscilloscope, it is always best to place several cycles of the voltage under investigation on the screen. The portrayal of one cycle is not generally satisfactory because of the distortion introduced by the part of the cycle that is *lost due to the retrace*. The presence of a single cycle on the screen would be satisfactory for examination purposes if the retrace time were ideal, that is, zero. But since this cannot be achieved, at least three cycles should be present on the screen. We mentioned this earlier in the text, but in this chapter we show the reason behind the statement.

In Fig. 10-103 are shown oscillograms made on two different oscilloscopes which differ widely in their characteristics. One is considered in the category of a television research oscilloscope, whereas the other is a maintenance instrument having reduced bandwidth and a lower frequency range for the time-base oscillator. In parts (A) and (B), the signal applied to the vertical deflection plates had a frequency of 40 kc. With three cycles on the screen, the sweep frequency was 40,000/3 or 13,333 cycles. These two periods correspond to 25 microseconds and 75 microseconds, respectively. In (A), that for the research oscilloscope, about 0.3 cycle is taken up by the retrace, so that the retrace period is 25 × 0.3, or 7.5 microseconds; for the maintenance oscilloscope (B), about 0.5 cycle is taken up by the retrace, which corresponds to approximately 12.5 microseconds. From the viewpoint of the portion of the sweep cycle occupied by the retrace, these two examples show about 10 per cent in the first case and about 17 per cent in the second instance.

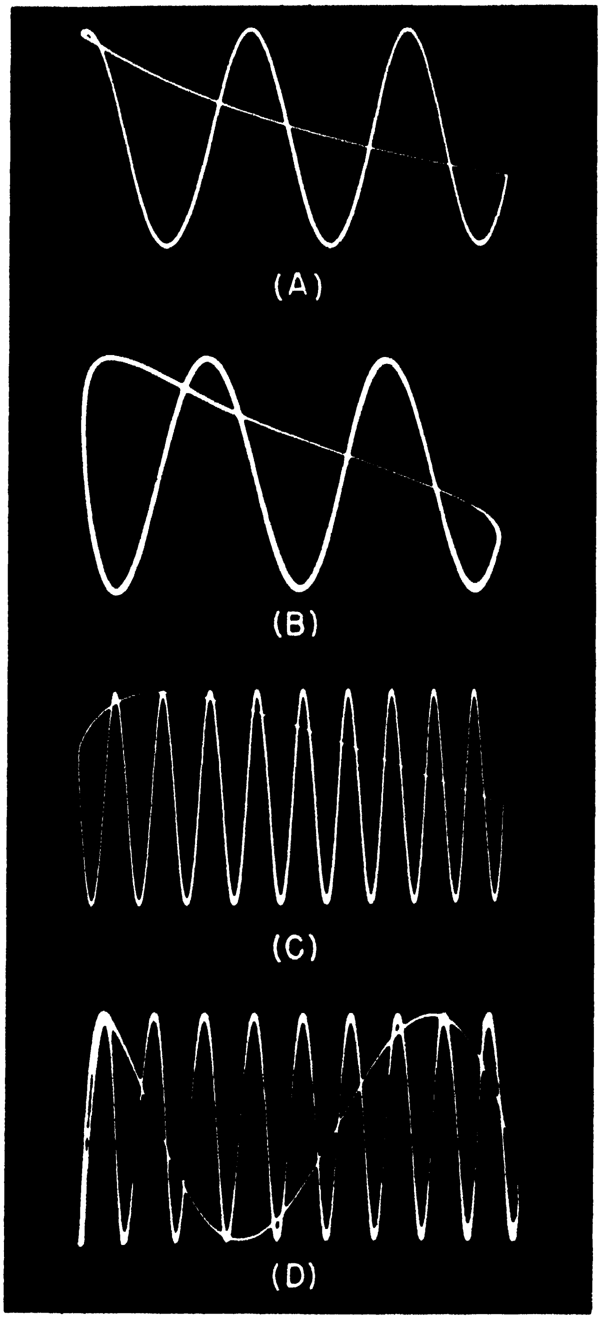


Fig. 10-103.—Illustrating the variation in the appearance of the return trace of a signal when displayed on different oscilloscopes. In (A) and (B) the signal has a trequency of 40 kc. In (C) and (D) a 70-kc signal is used.

A different condition develops on these two oscilloscopes when the signal and sweep frequencies are changed. In Fig. 10-103C and D are shown oscillograms for the research and maintenance instruments, respectively, with a 70-kc input signal. The sweep controls in each oscilloscope were set so that a like number of cycles are displayed on the screen. In each case 9 loops are visible on the upper part of the trace and 10 loops on the lower part of the trace. It might seem from this condition that both oscilloscope sweeps were adjusted to the same frequency, 7 kc. Actually that is not true. The sweep system shown in Fig. 10-103C was adjusted to the frequency as indicated by the nine primary loops and the return trace loop also located on the upper part of the display, thus making ten complete cycles. In Fig. 10-103D, however, there are ten primary loops and 1 return-trace loop on the lower part of the trace; on the upper part, there are nine primary loops and 2 return-trace loops, thus indicating the sweep frequency is 70/11, or 6.36 kc. Comparison of the sweep systems for frequency must therefore take the return trace into account.

Numerous conclusions can be drawn from this presentation. It is obvious that the time of the return trace in Fig. 10-103D is much greater than that of the return trace in Fig. 10-103C. The latter is about 1.5 cycles of the signal frequency, whereas the former is a small fraction of a cycle. Since a 70-kc wave has a cyclic period of about 14 microseconds, the return trace in one case is less than 7 microseconds, whereas in the other case it is more than 20 microseconds.

Oscillograms of the retrace-time period displayed herein lead to the conclusion that it decreases with the increase in sweep frequency, which is natural, since the higher the charging frequency in a repetitive system, the lower must be the discharge period in order to permit the increased speed of charging.

Visibility of the Retrace

The visibility of the retrace is determined by two conditions — its period and the portion of the cycle of the signal over which it is active. The velocity of the spot as it describes a cycle of the signal wave is a function of the frequency of the wave, but regardless of frequency, different parts of the cycle are traversed at different speeds. This is shown in general terms in Fig. 10-104 as applied to a sine wave. The point at which the spot is traveling fastest is where the greatest change in voltage with respect to time is taking place in the wave, this being the point where the sine wave passes through its zero value. After passing through zero, the rate of change gradually decreases and the velocity of the spot gradually decreases, until it is moving most slowly at the instants that the wave reaches its maximum amplitudes.

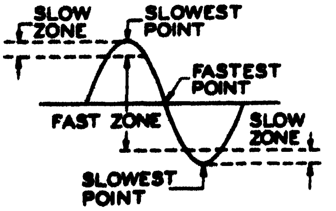


Fig. 10-104.—Illustrating spot speeds at different parts of a sine-wave trace.

This condition does not display any visible effects when the frequency is comparatively low, but when the signal frequency is very high, the effect of the different beam velocities is evident on the screen in the uneven brilliance of the trace. The trace is brightest at those points where the electron beam is moving across the screen most slowly, thus allowing maximum excitation of the screen material, and it is dimmest at those points where the beam is moving most rapidly. This explains why, in certain cases, only the peak amplitude points of a cycle of the wave are visible on the screen, these being the points which are traced by the slowly moving beam. An example of this is shown in Fig. 10-105.

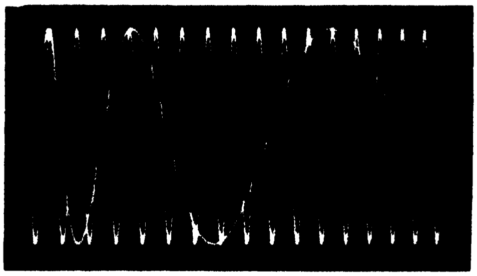


Fig. 10-105.—Showing different trace visibilities resulting from different spot speeds at different parts of a trace.

Such a condition is easily seen in the display of high-frequency square waves wherein the rise to maximum amplitude is too rapid to permit proper excitation of the screen. The result is that frequently only the constant amplitude points of the trace are visible as a series of parallel lines. This can be seen in the square-wave patterns which appeared earlier in this chapter.

Now referring once more to the return trace, it is understandable that if the return trace takes place on that portion of the signal cycle which corresponds to the greatest rate of change in amplitude, and the electron beam is moving most rapidly in the vertical direction, the return trace may not be visible as it completes its entire path. It may become visible at points where the electron beam is slowing down, but at the high-velocity points it may not be seen at all.

Concerning the visibility of the return trace, for that matter the complete cycle of a display, the greater the accelerating-anode potential, the greater will be the likelihood of a completely visible pattern. This explains why the higher the intended frequency of operation, the higher must be the accelerating-anode voltage in order to afford the required excitation of the screen material during the high-velocity beam excursions. These subjects relate to the *writing speed* of the electron beam. This is a separate subject and is discussed later in this chapter.

Blanking of the Retrace

The sweep retrace is not always visible on the screen. Fundamentally its purpose is to return the electron beam to its starting point for the next forward sweep, but this purpose does not require that it be present. Consequently, some oscilloscopes are designed so as to keep the return trace off the screen at all times. The return trace is blanked out by arranging that the *beam intensity* become so low during the retrace time as to be invisible on the screen.

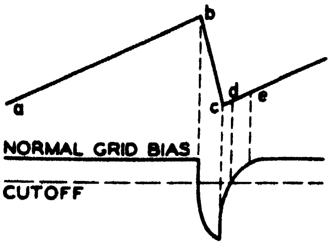


Fig. 10-106. — Coincidence of blanking and retrace voltages.

Return trace blanking is a facility which most oscilloscopes afford. In some instances, the related controls must be set to the *blanking* position; in other instances, blanking is a fixed condition in the device. The means of accomplishing blanking differs in different oscilloscopes, but in the main, two methods are used. One of these is the application of a momentary high negative voltage to the control grid of the cathode-ray tube, cutting off the electron beam. This conforms to the earlier reference to beam intensity or *Z*-axis modulation. By securing the blanking voltage from the sweep system and by proper shaping in a supplementary circuit (if one is used), the period of beam cutoff can be made to coincide with the period of the retrace, inclusive of whatever changes may occur in the retrace time as the sweep frequency is varied. The coincidence between these two actions is depicted in Fig. 10-106. The second method is equally effective and consists of making the cathode of the cathode-ray tube momentarily highly positive, by means of a positive pulse, which is the equivalent of making the control grid more negative.

The sawtooth voltage is passed through a differentiating circuit such as is shown in Fig. 10-107. The circuit components are *C* and *R*. C_b is very large compared to *C*, and serves only as an a-c bypass, placing the lower end of *R* at ground potential. During the linear rise in voltage *a-b* (see Fig. 10-106), the current through the resistor is constant and the resultant voltage drop is constant, although of such comparatively low value as to not alter the bias on the control grid. During the retrace period *b-c*, the rate of change of the current is very much more rapid and the discharge of *C* results in a high current through *R*, and therefore, a high voltage drop across that element. This drop is applied to the control grid in series with the regular grid bias and makes the grid very much more negative relative to the cathode, thus effectively cutting off the electron beam and extinguishing the spot on the screen.

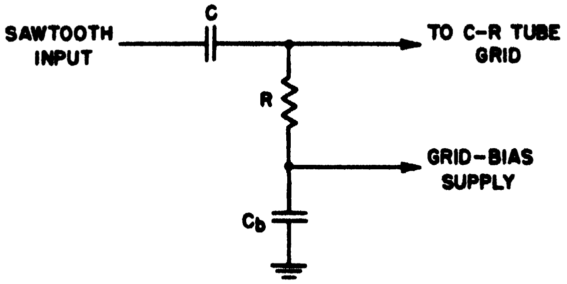


Fig. 10-107.—Differentiating circuit producing blanking voltage.

By adjusting the potentials properly, the additional bias necessary to cut off the beam is reached shortly after the start of the return trace, assuming that the change from the forward trace to the return retrace is abrupt as illustrated. The cutoff condition remains during the entire retrace period. In fact, in most instances, it remains there for a short period *after* the forward sweep has begun. This can be seen by noting the conditions of bias during the period *c-d* in the illustration. The presentation given may seem exaggerated, but in reality it is not so, because the action of blanking in virtually every practical case is such as to reduce the horizontal dimension of the trace on the screen. This does no harm because the horizontal amplitude control can be varied to give the desired trace length.

The reduction in trace length due to blanking is a function of the linearity of the sweep voltage. The more linear the sweep voltage during its cycle, and the shorter the flyback time of the sweep voltage, with the necessary abruptness at the maximum positive and maximum negative points, the less will be the reduction in horizontal length due to the blanking. On the other hand, and this might be said to be more frequently found at the higher sweep frequencies, the poorer the shape of the sweep voltage, that is, rounded corners at *b* and *c*, the greater the portion of the sweep which is made ineffective by the blanking action. With a badly distorted sweep voltage, the blanking action is capable of reducing the effective sweep length to about one-half of that which is normal.

The magnitude of blanking voltage necessary to accomplish the desired result is a function of the individual cathode-ray-tube operating parameters. Sometimes it is made variable by being passed through an amplifier whose gain can be controlled, and in many cases, the blanking circuit is a compromise between the limits set by the output of the sweep system at the lowest and highest frequencies. An arrangement of this type naturally is not as effective as the one wherein the amplitude of the blanking voltage may be controlled so as to afford the minimum effect on the trace length.

An interesting characteristic of blanking is shown in Fig. 10-106, between points *d* and *e* on the forward trace. Note that the cutoff voltage is reached at the point *d*, and that between *d* and *e* the control grid is becoming progressively less negative. It is only at *e* that the grid has reached its normal potential, so that the beam intensity and the spot luminosity increases *gradually* from *d* to *e*. This is evident in Fig. 10-108, which shows the difference in the appearance of the trace with and without blanking. Oscillogram (A) depicts a 70-kc wave with the sweep voltage adjusted to 5 kc, ratio of 14:1. Thirteen primary loops are visible at the bottom of the trace and the fourteenth loop appears on the return trace.

Fig. 10-108B is for the same conditions except that the return trace has been blanked; the fourteenth loop is missing and if an attempt were made to determine the frequency of the sweep relative to the frequency of the signal fed into the vertical plates, a misleading answer would be forthcoming. Now the ratio would *appear* to be 13:1, which is actually not the case. Therefore, if the sweep frequency is to be determined relative to a known frequency fed into the vertical plates, blanking should not be used. While this suggestion is general in its application, we recognize that in some oscilloscopes the return trace has an extremely short period and blanking would not exceed a fraction of a single cycle, therefore, identification of frequency ratios could be accomplished even with blanking. Nevertheless, greatest ease of computation is attained when blanking is removed.

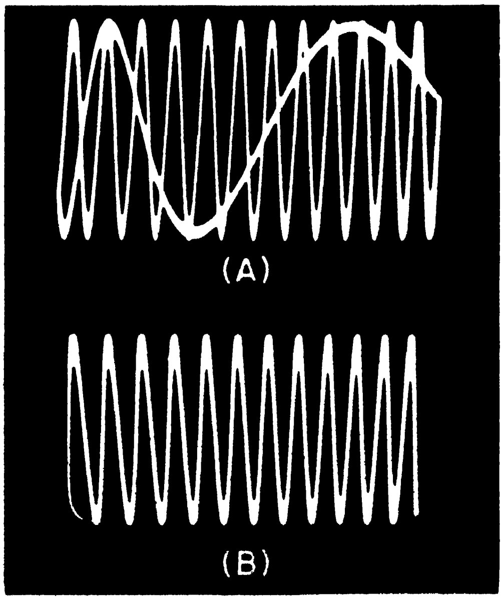


Fig. 10-108.—Typical sine wave, (A) without blanking and (B) with blanking.

It may appear from Fig. 10-107 and the discussion which accompanied it that blanking of the trace is limited to the sweep retrace. Such is definitely not the case. Any portion of the trace may be blanked out by appropriate means, which simply means that, at the proper instant, the control grid is made sufficiently negative to extinguish the beam. For example, in Fig. 10-109 are shown three oscillograms illustrating

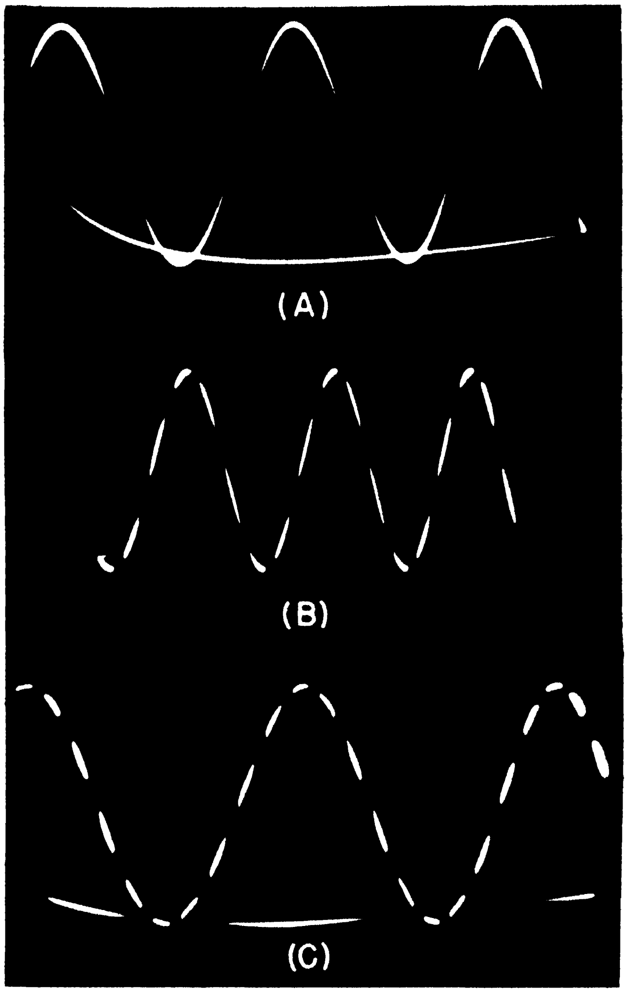


Fig. 10-109.—Intensity modulation, (A) with modulation frequency twice the signal frequency, (B) with modulation frequency six times the signal frequency, and (C) with modulation frequency ten times the signal frequency.

blanking or intensity modulation although the purpose of the action may be entirely different from the reason for blanking the retrace. At the moment the reason is secondary to the means, although it can be said that sometimes some sort of marker indications representing time or some other quantity may be desired on the trace, and blanking is one of the two means of placing such markers on the trace.

Oscillogram (A) depicts a sine wave of voltage applied to the deflection plates of a cathode-ray tube when the sweep frequency is one-third of that of the deflection signal and a sine wave of voltage is applied to the control grid of the cathode-ray tube for blanking the beam twice during each deflection-voltage cycle. This means that the blanking signal frequency is twice the deflection-signal frequency. Since the grid is made more negative than normal in order to achieve blanking (or the cathode is made more positive), it is the negative peaks of the blanking signal which are effective in creating this type of pattern.

Two other illustrations of blanking are shown in Fig. 10-109 B and C. In the former case, the blanking signal has a frequency equal to six times the deflection-voltage frequency. This is indicated by the presence of six blank spaces in each cycle of the trace. Oscillogram (C) displays a 10:1 frequency relationship between the blanking signal and the deflection voltage. In both illustrations, the blanking voltage is a sine wave and the pattern was developed by utilizing the negative peaks to drive the control-grid potential beyond cutoff.

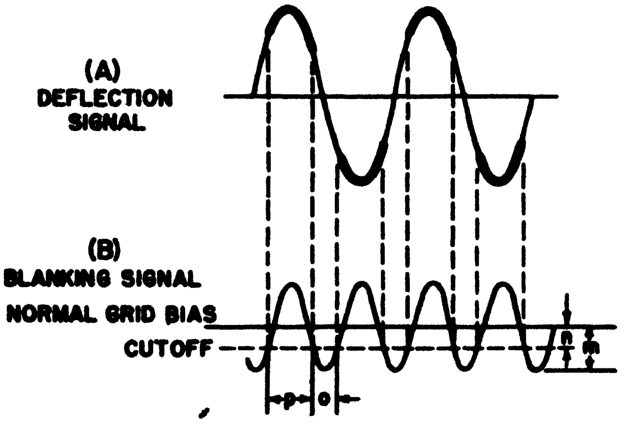


Fig. 10-110.—Blanking and signal voltage relations occurring in Fig. 10-109(A).

The period of blanking, or the period during which the beam is extinguished, occurs twice during each cycle of the deflection signal in (A) because two negative peaks of the blanking signal occur during each cycle of the deflection voltage. The time intervals of these periods determine the portion of the deflection signal which is blanked out. This is illustrated in Fig. 10-110, wherein are shown the vertical-deflection signal and the blanking signal, as well as the control-grid potential. The starting phase of these two signals is such that the blanking signal leads the deflection voltage by 45°. If you compare the heavy lines of curve (A) with Fig. 10-109A, you will note the correspondence between the blanked portions of the two traces. You can readily see that a difference in starting phase between the deflection signal and the blanking signal can change the portion of the cycle which will be blanked. Another controlling factor is the relative level of the grid potential and the cutoff voltage with respect to the amplitude of the blanking signal. The higher the peak value *m* of the blanking signal relative to the cutoff voltage *n*, the greater will be the portion of deflection signal which will be blanked. This can be visualized by noting the projection of the dimension *o* onto the deflection signal. The greater the ratio of *m* to *n*, the wider the dimension *o* and the greater the time during which the blanking signal keeps the grid cut off.

An interesting condition is evident in Fig. 10-109B and C. You can see that the segments of the trace which describe the cycle are not uniform. This stems from the fact that the velocity of the electron beam is a function of the rate of change of deflection voltage. The greatest rate of change takes place each side of the zero axis and it is here that the velocity of the deflected beam is greatest; it is slowest at the top. Since the horizontal displacement is linear throughout the cycle, movement of the spot will be greatest where the rate of change is greatest and will be least where the rate of change is the lowest. Hence the trace lengths on the slopes of the cycle are longer than the trace around the peaks.

Another interesting condition which appears in Fig. 10-109 is that relating to the return trace. In (B) of this figure, the return trace has been blanked, which means that the blanking action at the grid can be of widely different time intervals depending entirely upon the operating parameters of the circuits which feed the controlling signals to the control grid. In this case, the sweep-retrace blanking signal came from one source and was applied to the grid, and the other blanking signal came from an entirely different source and was applied to the same control grid.

In Fig. 10-109C the retrace was permitted to remain and it is evident that the sine-wave blanking signal acted on the retrace as well as on the forward trace. This is indicated by the gaps in the retrace pattern. Knowing that the sine-wave blanking signal blanked the vertical-deflection signal ten times in each cycle,

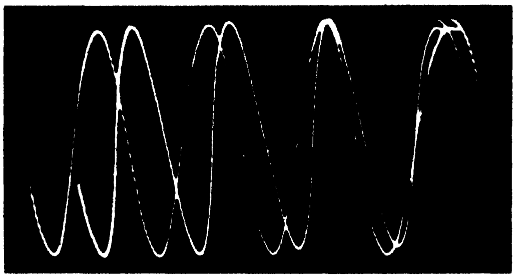


Fig. 10-111.—Distinguishing between traces in a dual-beam tube by intensity modulating one of them.

the presence of four blank spaces in the retrace indicates that the period of the retrace was equal to four-tenths of the period of one cycle of the vertical signal.

The use of such a sine-wave blanking signal is one means of distinguishing between two traces which appear simultaneously on the face of a dual-beam tube. These two traces may indicate actions in different parts of a circuit, current or voltage relationships, or any other two quantities. By applying a sine-wave blanking voltage with a frequency which is high relative to the frequency of the voltage applied to the vertical-deflection plates, the gaps in the trace can be made so small as to show no visible interruptions, yet enable instantaneous identification of which display is which. A typical example of this is shown in Fig. 10-111.

Markers by Blanking

The two kinds of blanking signals shown so far are not representative of the full range of varieties which may be employed or encountered. For example, markers may be placed on a time-base trace so as to check its linearity by making the blanking voltage a series of sharp pulses of short duration and of a specific repetition frequency. With the frequency of the blanking pulse known, the time interval between blanks in the trace can be calculated easily. If the sweep trace is linear throughout its length, the blank spots will be equidistant from each other as shown in Fig. 10-112. If the separation between them is not equal, the degree of nonlinearity can be determined. Such pulses can be developed by differentiating a square wave and clipping either the positive or negative portion, depending on whether the blanking is applied to the control grid or to the cathode.

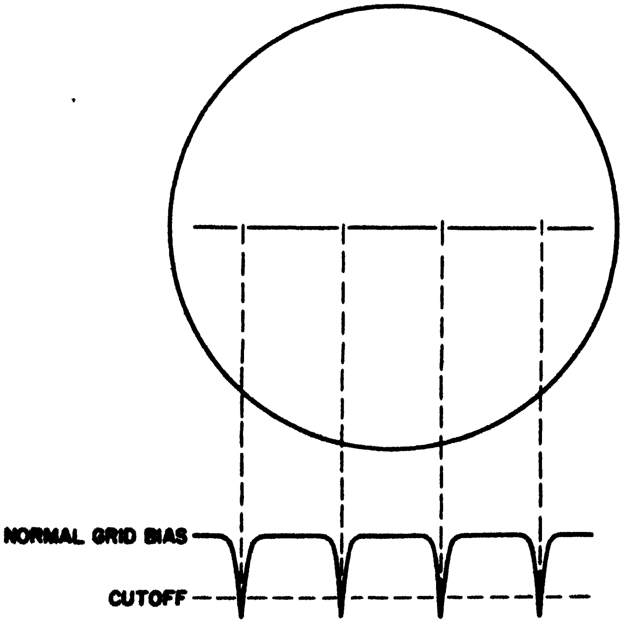


Fig. 10-112.—Checking the linearity of a trace by timing markers.

Markers by Intensity Modulation

There are some who insist that blanking and intensity modulation are one and the same thing. Personally, we do not agree, simply because we consider the beam as the basis of the comparison. By definition, intensity modulation is the process of conveying some intelligence by varying the intensity of the beam; the difference in intensity being the means of obtaining the intelligence.

The practical difference between blanking and intensity modulation *as we see it* is that in the former case the normal operation of the cathode-ray tube demands that the electron beam be active and a trace be initially present on the beam. Any modulation of this beam by decreasing its intensity, that is, extinguishing the beam, is blanking. The second process, which we consider intensity modulation, is operation such that the electron beam is normally *extinguished;* when modulation is applied, the trace *appears* on the screen. The extent to which the beam is extinguished does not matter; it can be completely cut off or its intensity reduced to where it is just barely visible. In either case, the basic conditions for blanking and intensity modulation are different. In the former, the trace is removed from the screen by the control signal, and in the latter, it is put on the screen by the control signal. In blanking, the control grid is driven to and beyond cutoff; in intensity modulation, the control grid normally is at or beyond cutoff and is driven above cutoff.

The appearance of the trace on the screen during intensity modulation can be the same as for blanking. The choice of one process over the other process of developing a trace is one of purpose, convenience, economy of operation, cathode-ray-tube life, and other related details. It is understandable that a screen without any signal display on it is easier on the eye for a period of observation than one on which traces are present and intelligence appears when they are removed. It is more effective to convey intelligence by making something which has been absent appear on a screen. Hence the use of intensity modulation in numerous applications which involve short signal duration conditions, or long hours of observation. Many radar displays make use of intensity modulation.

Patterns such as those shown in Fig. 10-109 can just as readily be duplicated by intensity modulation as by blanking. All that would be necessary is to change the normal operating point of the control grid so that by virtue of the bias applied, the normal operating condition is the beam extinguished.

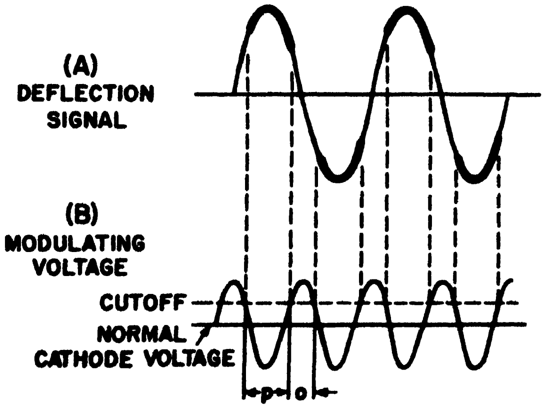


Fig. 10-113.—Modulation and signal voltage relations necessary to produce the trace of Fig. 10-109(A).

The development of a trace, similar to that shown in Fig. 10-109A, by intensity modulation applied to the cathode would be done as shown in Fig. 10-113. Here the normal potential of the cathode without any modulating signal input is sufficiently positive relative to the control grid that the latter is beyond cutoff and keeps the beam extinguished. The modulating voltage is applied to the cathode and its polarity is such that, for the period *p*, it applies a gradually increasing negative potential to the cathode, thus driving the control grid above cutoff and permitting the beam to exist. The beam, therefore, would trace out the heavy-line pattern. Then as the modulating voltage rises in the positive direction, the beam will again be extinguished and remain that way during the period *o*. If the modulating signal were removed entirely, the beam would remain extinguished since the normal operating potential applied between the cathode and control grid is such, in this example, as to keep the beam cut off.

The process of marking by *increasing* the intensity of the trace at certain prescribed instants is accomplished by increasing the beam density. Either the cathode or control-grid voltage is varied in such manner as to increase the number of electrons in the beam.

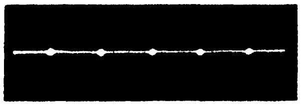


Fig. 10-114.—Intensity modulation by increasing the trace intensity.

An example of this is shown in Fig. 10-114. Here a conventional sweep trace is intensified at five intervals. It is a means of checking linearity of the sweep system, but the technique may serve any number of purposes of identification. The apparent increase in line width at the moments of beam intensification is due to two effects. One of these is associated with the conditions set forth while making this illustration. The intensity of the trace prior to intensification was purposely kept below normal so as to show most clearly the increase in beam intensity as the marker signals were applied. The other is the general condition that intensification does produce a beam of greater diameter, especially when the intensity and focus controls are preset for optimum beam dimensions.

A combination of the electrical conditions described in connection with Figs. 10-113 and 10-114 might be said to be the basis of producing such patterns as were shown in Fig. 1-9D, the radar screen which depicted the Hudson River near New York City. Each dot of light on that screen is a reflection from the target and is applied to the cathode-ray tube in such manner as to intensify the beam, which otherwise is either extinguished or if it is producing a circular scale of distance from the screen to the target, is of relatively low intensity.

As in the case of blanking, the shape of the voltage which may be used for intensity modulation varies, being determined by the manner in which the beam is to be turned on and off, and for what periods these states are to prevail. The use of sine-wave intensity modulation is a very simple example and not necessarily typical of the major application of intensity modulation. Admittedly, the facility is not generally available in the simplest of oscilloscopes. However, as in the case of blanking, provision for intensity modulation is to be found in numerous oscilloscopes, which are not too different from the basic variety.

Triggered Sweeps

The two examples of time-base voltages discussed so far were of the repetitive character. The fact that their waveforms were different, one being sinusoidal and the other sawtooth, does not matter. These voltages repeat themselves as long as the operating voltages are applied to the circuits which produce them. Some oscilloscopes afford an additional time-base voltage; it is sawtooth in shape but it does not repeat itself. One might say that it is a "single-shot" sweep which depends upon a "trigger" voltage or signal to set it into operation. Such time-base systems are known by a number of names such as "triggered sweep," "nonrepetitive sweep," "single-shot sweep" — all of which mean the same thing.

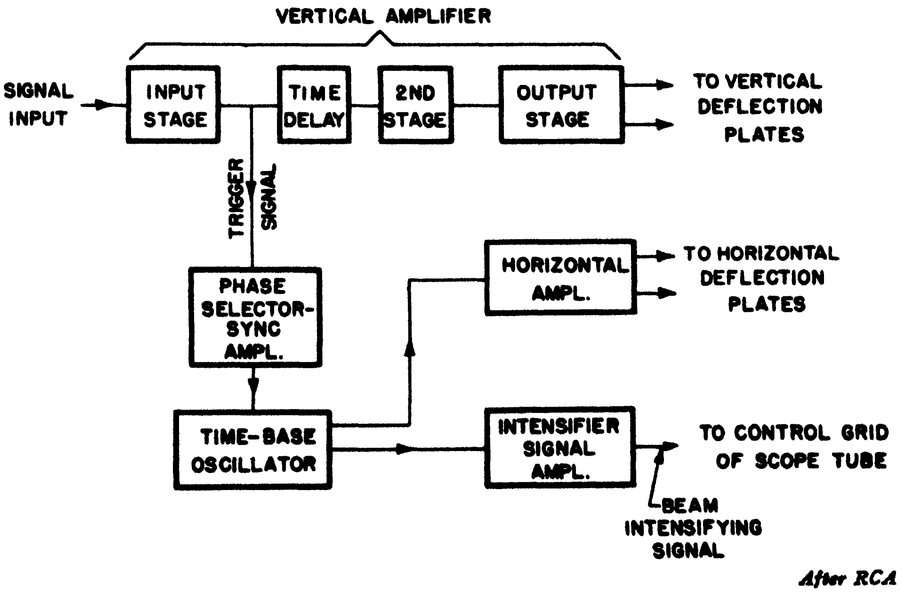
Their reason for being is the need for observing extremely rapid pulses or transients which may reappear at uniform intervals or which may have an irregular repetition rate. In either case, it is difficult to observe such transients with the normal sawtooth sweep because, as a rule, it is too slow in its action. Moreover, if the sweep is repetitive and the signal is not, the timing between the two becomes a problem, especially if the sweep must be expanded so as to enable close study of the transient. In addition, synchronization is virtually impossible between an irregularly appearing transient and a conventional repetitive sweep.

To solve all of these problems, especially the need for stationary pattern for study purposes, the triggered sweep was conceived. As a matter of fact, a certain classification of oscilloscopes capable of such operations came into being during World War II. This is the device identified as a "synchroscope," which might be defined as a synchronized oscilloscope. All oscilloscopes provide for synchronization between the signal to be observed and the time-base-voltage system; however, the synchroscope, being specifically designed for such work, differs from the conventional instrument in that it does not afford the other capabilities found in the conventional research or maintenance cathode-ray oscilloscope.

Recognizing the expanded application of such short-duration pulses in many fields of activity, the manufacturers of conventional oscilloscopes added these nonrepetitive sweep systems to some of their products, in the form of separate and distinct systems, or as elaborations on the repetitive system whereby a single-cycle sweep of any frequency within its range becomes available. It is not yet universal, being limited in a sense to the more expensive research devices. One of the primary conditions of this system involves the synchronization between the transient and the sweep. Both being of very short duration and usually of irregular repetition rate, the display of the signal demands that the sweep trace begin its scanning movement at exactly the same instant as the signal begins its amplitude variations, and that it remains in action as long as the signal is present. Being of very short duration and perhaps appearing only at irregular intervals, it stands to reason that the utmost in synchronization is desirable in order to show the complete signal. If the sweep starts even a fraction of a microsecond later, then the signal begins its amplitude variations, a portion of the vertical signal will be lost in the display and the ability to study it will be impaired.

The Delay Circuit

By all practical everyday methods of reasoning, a fraction of a microsecond is a negligible period of time, but when the problem is the observation of a transient



After RCA

Fig. 10-115.—Block diagram of an oscilloscope incorporating a time-delay circuit.

which lasts only one microsecond or less, fractions of this time interval have much meaning. In order to accomplish the desired synchronization, two methods are used in oscilloscopes. One of these consists of a system which delays the signal fed into the vertical amplifier so that it reaches the vertical-deflection plates about 0.2 microsecond later than the start of the sweep voltage which is fed to the horizontal-deflection plates. The *delay circuit* is an integral part of the vertical amplifier; for other applications of the oscilloscope, the short time delay means nothing. It plays its important role in the observation of transients when the sweep is triggered into action by the applied signal, from a point which is ahead of the delay line. This assures the start of the sweep trace 0.2 microsecond ahead of the appearance of the signal at the vertical plates. The organization of this system is shown in simplified form, with all switching removed, in the block diagram of Fig. 10-115.

Another very interesting condition is associated with the general use of triggered sweeps. While it is applicable to the system shown in Fig. 10-115, it should be understood that it is to be found in other organizations of such oscilloscopes. This is important because time-delay networks are not standard in all oscilloscopes which provide the facility of high-speed-pulse observation. What we have in mind is the action of the system identified with the intensification of the trace.

Since the signal and sweep are synchronized in action, the sweep voltage is not applied to the horizontal-deflection plates via the horizontal amplifier until the sweep generator has been triggered by a signal from the first stage of the vertical amplifier. Therefore, the screen of the tube displays only the tiny spot where the undeflected beam strikes the screen. Since velocity of the beam is very great as it traverses the screen, and since the intensity of the trace on the screen is a function of the amount of time allowed the beam electrons to excite the screen (as well as the accelerating voltage, which is fixed in value), a supplementary *positive* voltage is applied to the cathode-ray-tube grid simultaneously with the application of the sweep voltage. This voltage intensifies the trace so as to make it more brilliant and easier to see. As is evident by the block diagram, the intensifying voltage is synchronized with the sweep voltage and the intensifier amplifier receives its impulse from the sweep source, therefore, the intensifying voltage is applied as long as the sweep voltage is applied. In a sense, you can consider this intensity modulation of the beam, for its intensity is increased above normal as a consequence of the sweep signal, if not the vertical-deflection signal.

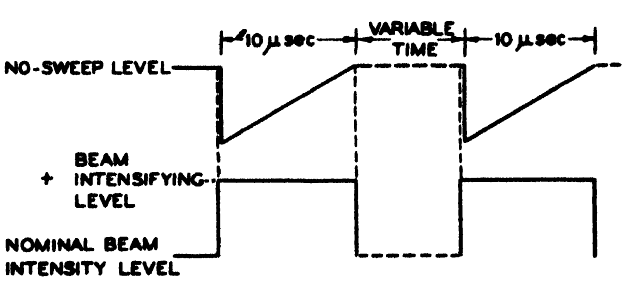


Fig. 10-116.—Voltage relationships during the period of operation of a triggered sweep using beam intensification.

This action is illustrated in Fig. 10-116. Two periods are shown for the sawtooth sweep, that which is the condition of no sweep and that which represents the sudden discharge and the charging period conforming to the forward trace. It is this forward trace which is timed with the square-wave intensifier voltage. In view of the inability to achieve a zero time-discharge condition, the instant of beam intensification is slightly ahead of the start of the sweep period; the return to normal beam intensity coincides with the end of the sweep period.

The somewhat different than normal shape of the sweep trace justifies a slight comment. Actually the sweep is no different than for the repetitive sweep; it is only the circuit conditions which are different. In order that the sweep be ready for action, the capacitive element is kept charged while awaiting the trigger pulse. This pulse actually *discharges* the capacitor so as to start the sweep cycle. What we show as the "no-sweep level" actually is the stable state of the sweep with the capacitor fully charged. Once the sweep cycle has been completed, the time-base oscillator remains in the stable condition while awaiting the next trigger pulse.

The beam-intensifying pulse shown in Fig. 10-116 should not be confused with the trigger pulse. The latter originates in the block identified as "Phase Selector-Sync Amp." This leads to a possible question: Is the duration of the triggered sweep always the same as the duration of the signal?" The answer depends on the time duration of the signal and the *period* of the sweep or the sweep frequency which is selected by means of a selector switch. Different oscilloscopes having triggered sweeps afford different means of controlling the duration of the sweep. In the specific oscilloscope being discussed (RCA WO-79A) and which is described in detail in Chapter 22, the time duration of the sweep is determined by the position of the time-base-oscillator frequency controls. These are the same controls which permit the selection of the sweep frequency for repetitive operation, except that when in the "triggered" position, only a single cycle of the sweep is applied to the horizontal-deflection system.

In other oscilloscopes, which also are described in Chapter 22, the triggered sweeps are of fixed time duration and are selected by means of a special switch that is apart from the repetitive sweep-frequency control. It is always best if the period of the triggered sweep is just slightly longer than the signal period. When this happens, the signal display is the largest, since it is spread to almost the full length of the sweep trace.

Trigger Generators

A refinement which is found in some oscilloscopes, especially those intended for research rather than maintenance, is the trigger generator. This is a multivibrator system capable of operating over a frequency range and which produces a short-duration trigger pulse of either positive or negative character. The purpose of this generator is to enable the starting of circuits whose output waveform must be examined by means of the triggered time base. Through suitable circuits, the trigger generator which starts the source of the wave to be examined can be timed relative to the start of the triggered sweep so that they begin in unison; or the start of the sweep may lead or lag the signal source by any prescribed amount up to perhaps 1,000 microseconds.

When this facility is available, the triggered sweep may be started in such time relation to the signal being examined, as to allow observation of any portion of the signal cycle — the leading edge, the lagging edge, or any other portion of the wave.

This is illustrated in Fig. 10-117. Let us assume that (A) shows the shape of the voltage which is fed into the vertical-deflection system. Curves (B), (C), and (D) show the sweep voltage from the single-cycle sweep. By means of the trigger generator, the start of the sweep can be made to coincide with the beginning of the signal voltage, as shown in part (B). If desired, the sweep can be started after the time lapse *a-b* so that the beginning of the sweep *b'* is just ahead of the linear rise; this appears in part (C). Or, as in (D), the sweep can be started after the time lapse *a-c,* so that the sweep starting point *c'* will be just ahead of the steep rise in the vertical signal. The capabilities of such devices are tremendous, but understandably they are required only by those who do work of this kind. The average individual who employs the oscilloscope for the everyday maintenance operations cannot justify the expenditure for oscilloscopes of this kind, even though the maintenance requirements are more than satisfied by these elaborate devices.

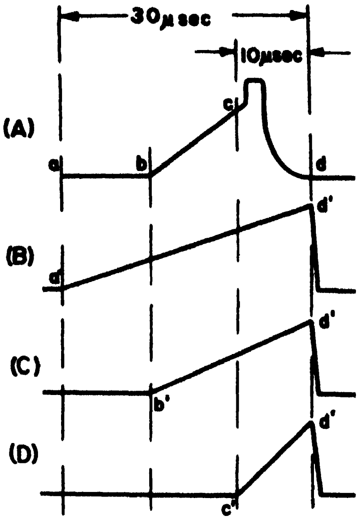


Fig. 10-117.—(A) Signal under observation, (B) normal triggered sweep length, (C) sweep shortened to accommodate the entire signal, and (D) sweep shortened to display only the interesting portion of the signal.

The advantage offered by such a system lies in the ability to utilize the full sweep width for the portrayal of the portion of the signal wave that is of greatest interest. This may not be clearly evident in Fig. 10-117, but if it is understood that the dimensions *a'-d', b'-d',* and *c'-d'* along each sweep represent identical widths of the trace on the screen, then the expanded display shown in Fig. 10-118 will be understood.

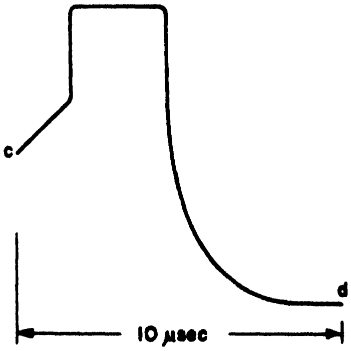


Fig. 10-118.—Enlarged display resulting from Fig. 10-117(D).

By starting the triggered sweep of Fig. 10-117D at the proper time and with a period equal to ⅓ of Fig. 10-117A, the 10-microsecond portion of the vertical signal was expanded to the same horizontal dimension as the original, thus making the characteristics of the wave much easier to study.

Sweep Speed

In the normal course of oscilloscope operation, little concern is felt for the speed with which the electron beam completes its horizontal excursion across the screen. The sweep is set at a certain frequency and the fact that the time-base trace of a certain dimension corresponds to a certain frequency is taken at face value. Sometimes it is necessary to determine the duration of an action or the duration of a part of an action. Knowledge of just the frequency of the sweep is not sufficient to enable such measurement. It then becomes necessary to know the speed with which the beam completes its entire path or fractions of that path length. When that is known, simple lineal measurement of the distance between points on the trace is interpretable in time duration. Some oscilloscopes provide this facility in the form of calibrated sweep speeds.

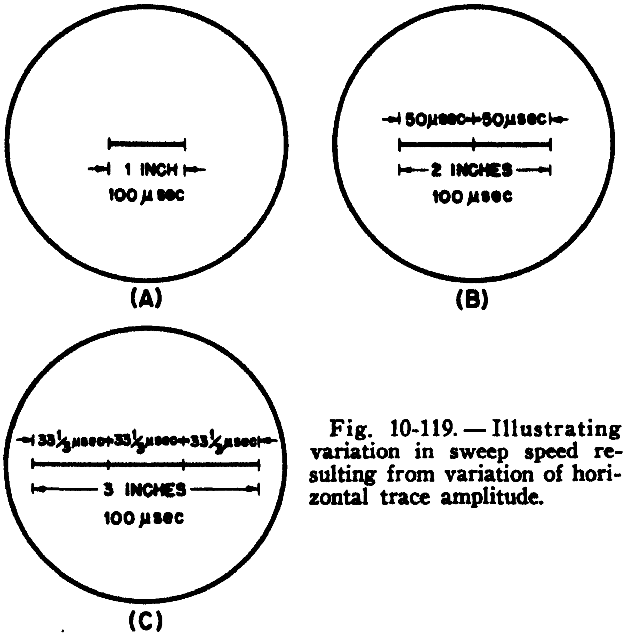


Fig. 10-119. — Illustrating variation in sweep speed resulting from variation of horizontal trace amplitude.

For example, let us consider the three screen faces in Fig. 10-119. The frequency of the sweep which produces the horizontal trace in all three cases is the same, arbitrarily 10,000 cps. This corresponds to a cyclic period of 100 μsec, in which case we are neglecting the retrace portion of the sweep. Let us further imagine, as shown in illustration (A), that the length of the sweep trace is 1 inch. Since the period of a single cycle is 100 μsec, the time required for the beam to complete a single horizontal excursion is 100 μsec.

Now we increase the sweep *amplitude* so that the trace is 2 inches long as illustrated in (B), instead of the previous 1 inch in (A). The frequency of the sweep is still the same 10,000 cps. Expanding the horizontal dimension of the trace for the same sweep frequency setting simply means a high deflection voltage and this means greater speed as the beam is moving under the influence of the deflection fields. (See Chapter 2.) It will complete the longer trace in exactly the same over-all time required for the shorter trace, but since it must cover twice the distance than it did before, it must be moving at twice the velocity, so that the sweep speed for 1 inch of distance now is 50 μsec.

If we increase the horizontal trace dimension even more, say to 3 inches, as shown in illustration (C), it means a still greater velocity during deflection so that the entire trace will be completed in the original 100 μsec. Since the over-all distance has been increased threefold, the speed per unit distance (in this case 1 inch) must be increased threefold, so that 1 inch is completed in $^{100}/_{3}$ or 33⅓ μsec.

Now if we set a standard trace dimension as the basis, it is possible to indicate the sweep speed in terms of time relative to the frequency. This is done in oscilloscope manufacturers' specifications. Suppose that we assume a standard trace length of 4 inches. Whatever may be the length mentioned in the specifications, it is that dimension which should afford a linear sweep. We assume this to be the length stated, although in the final analysis, it is determined by the screen diameter. For a 3-inch screen, it may be 2 inches and for a 5-inch screen, it may be 4 inches.

If the rated trace dimension is 4 inches, then sweep speeds might be stated as follows:

Sweep Frequency	*Sweep Speed, per inch*
1.0 Mc	0.25 μsec
250 kc	1.0 μsec
100 kc	2.5 μsec
50 kc	5.0 μsec
10 kc	25.0 μsec
1.0 kc	250.0 μsec

The measurement of time duration of an action shown by such sweeps is simply a matter of determining the space along the horizontal dimension which is occupied by the action.

Sweep Calibrator

Some oscilloscopes provide the means for calibrating the sweep speed by means of accurately calibrated frequencies which are used to supply the vertical deflection. The time-base speed then is determined by measuring the distance along the horizontal dimension which includes one or more cycles of the known frequency. Thus if a 1-Mc voltage is applied to the vertical-deflection plates and 2 cycles are shown on the screen when the horizontal amplitude is 2 inches, the sweep speed is 1 μsec per inch. Obviously, such calibration is a function of the horizontal dimension as well as the frequency of the sweep voltage, for if 4 cycles of the 1-Mc vertical signal are shown on the screen and the horizontal trace is 2 inches long, the sweep speed is 4 μsec for the entire length and the speed per inch is 2 μsec.

Vertical Amplitude Calibrators

Another facility which is contained in some oscilloscopes is the means for measuring the peak-to-peak amplitude of the voltage applied to the vertical-deflection system. This is known by different names, foremost of which is simply calibrator. The system furnishes a variable square-wave voltage of known amplitude which may be applied to the cathode-ray tube via the vertical amplifier in place of the signal being measured. When the amplitude of the calibrated voltage trace corresponds to the amplitude of the signal under investigation, the calibrating voltage is read off the associated scale marked on the panel and it identifies the amplitude of the signal at the input of the vertical amplifier.

Such voltage calibrator systems are available separately as an oscilloscope accessory and are described fully in Chapter 14. Their application is described in Chapter 19.

POWER-SUPPLY CIRCUITS AND OTHER CONSIDERATIONS

Of the many sections of the oscilloscope, the power supply is the most commonplace. It is not necessarily the simplest, but it is the section which, while indispensable to the proper functioning of the device, contributes little to its performance specifications. Nevertheless, it too bears certain ratings and these have an effect on its suitability for application in certain locations. These are the frequency and voltage ratings of the power-supply transformer.

Power-Supply Frequency

Every power transformer bears two general ratings, namely, frequency and voltage. Both are related to the constants of the power system at the point where the device is to be used. As far as the North American continent is concerned all but one nation makes use of 60 cps. The exception is the Panama Canal Zone where only 25-cps supply is available. However, even in nations where a 60-cps supply is used predominantly, some areas may use a 50- or 25-cps supply.

As far as the rest of the world is concerned, the frequencies of the power-supply systems range from 25 to 60 cps. The electrical industry has grouped these frequencies into two general categories, 25 to 40 cps and 50 to 60 cps, and have designed transformers accordingly. Generally speaking, transformers which bear 50- to 60-cps ratings are *not* suitable for use on 25- to 40-cps power lines. This immediately establishes the limitation that an oscilloscope which contains a power transformer designed for 50 to 60 cps is not suitable for use on power-supply systems rated between 25 and 40 cps.

On the other hand, power transformers which are rated at 25 to 40 cps *may* be used on 50- to 60-cps power lines, provided that the necessary correction is made in the turns ratios so as to supply the output voltages needed by the circuits in the oscilloscope. Such an arrangement may be found in equipment intended to have universal utility. The transformers are designed for 25- to 40-cps power lines and contain the means which permit its application to the higher-frequency systems.

Power-Supply Input Voltage

The line voltage or power-supply input voltage rating is so closely tied to the power-supply frequency that it should not be dealt with separately. This factor does not exert the same degree of control as does the frequency. While it is true that a wide variety of voltages are used in the power systems of the world, the adaption of one value to another is comparatively simple. However, it must be understood that certain limitations do exist in use.

In the United States, a-c power lines generally provide from 110 to 120 volts. In a few instances, the local supply voltage exceeds this higher limit, being from 125 to perhaps 130 volts. Most equipment built in the United States and designed for operation on a-c lines, uses 117 volts as the standard line voltage. This is a compromise value.

When the power transformer is tapped for different line voltages, changing from one to another according to the existing need, poses no problem; it may be the manipulation of a switch inside the device or it may be the changing of one lead. Care must be exercised so as not to use a transformer intended for 150- or 220-volt operation on a 110- to 125-volt line, because the oscilloscope will function very poorly, if at all. On the other hand, connecting a 110- to 125-volt transformer to a 150- or 220-volt circuit will result in damage to the instrument because of the resultant increase in secondary voltages.

Where a difference in voltage rating of the power transformer and the local power supply exists, *assuming that the frequency is correct* and automatic internal means of adoption do not exist, the remedy is simple. The most convenient arrangement is the use of either a step-down or a step-up transformer, whichever is required, between the power line and the input of the power transformer of the oscilloscope so as to supply the proper voltage to the device. Thus if the device is rated at 110 to 125 volts and the local power-line voltage is 220 volts, a step-down transformer changing 220 volts a-c to 110 volts is required, as shown in Fig. 10-120. The other values are handled in like manner, except for the direction of voltage transformation, whether up or down, and the specific values involved.

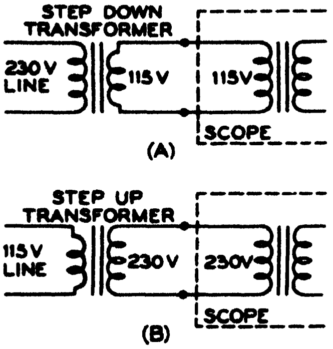


Fig. 10-120.—Step-up and step-down oscilloscope power transformers.

Where rigid control of a fluctuating line voltage is desired, voltage-regulator transformers may be used between the power-line circuit and the oscilloscope power transformer. These transformers tend to maintain the voltage at the input of the oscilloscope power transformer constant over a range of power-line voltages, but even this does not apply over a limitless range. Each regulator transformer bears a rating of input voltages over which it is effective. Moreover, every such transformer maintains its efficiency only when it is feeding a load which does not exceed its rating. Therefore, when a voltage-regulating transformer is utilized to feed the power transformer in the oscilloscope, its power handling rating must be *at least equal* to the power rating of the oscilloscope. If at all possible, the rating of the regulating transformer should be from 15 to 20 per cent above the load that it must serve.

While on the subject of voltage-regulating transformers, we might voice a cautionary note concerning the waveform of the output voltage from these devices. Unless specifically designed to avoid doing so, these units tend to distort the wave very badly, and will interfere with the performance of the oscilloscope which is connected to it. Therefore, the purchaser of such a device should make certain that the waveform of the output voltage contains a minimum of harmonics. In fact some manufacturers of voltage-regulating transformers provide filters for this purpose.

Power-Supply Circuit Arrangements

In view of the contents of Chapter 22 in which are shown and discussed the complete circuits of oscilloscopes, we shall treat here only the fundamentals of power supplies and the variations from general circuit format. Power supplies used in oscilloscopes are no different from systems used for like functions in other kinds of electronic equipment. If any significant differences exist between them they are found in the power transformers and the arrangements for polarity and grounding.

Relative to the latter, the oscilloscope differs from the majority of electronic instruments which derive their operating voltages from an a-c operated power supply. In the majority of electronic circuits, the negative wiring bus is grounded. This is true in most of the sections of the oscilloscope too, but a variation is introduced in the manner in which the accelerating potentials applied to the cathode-ray tube are treated. Unlike the usual arrangement, the accelerating anode of the cathode-ray tube is grounded, although it is subject to a relatively high positive potential. This is done in a high-voltage power supply by grounding the positive output terminal. In that way the accelerating anode and the deflection plates are at substantially the same mean potential and this tends to minimize defocusing of the beam during deflection. This particular subject was treated in Chapters 5 and 6.

Because of this situation, special types of circuit arrangements are used in the power-supply section in order that the high positive voltage required by the accelerating anode be grounded yet not interfere with the low-voltage systems of the power supply.

Returning again to the purpose of the power supply, it is simple to say that it is the source of the operating voltages. That is correct up to a point. It is a constant d-c voltage within certain limitations. Every power-supply system is rated in d-c voltage output and d-c output current. The two ratings are interdependent, and indicate the voltage output at a specified current drain. In the usual run of power supplies, an *increase* in *current* drain means a *decrease* in output *voltage*, but even this cannot be carried too far, for if the current drain is excessive, that is, exceeds the limits of the design, the output voltage becomes subject to major fluctuations. These occur at a frequency which will equal the power line frequency or twice this frequency, depending upon whether the unit is a half-wave or a full-wave rectifier. Restated, excessive current drain not only reduces the output voltage but creates a "hum" problem. This is the case in the *unregulated* power supply.

The *regulated* power supply contains a system by means of which the voltage output is maintained constant, within certain limits, over a range of current drain. This type, like the unregulated, is used in oscilloscopes, but the controlled variety feeds those tubes in d-c amplifiers which demand a constant voltage over a range of plate currents. Even this type has its limitations; if the stated current drain limits are exceeded, the regulation will fail and it will behave just like an unregulated supply.

Component Parts of Power Supplies

All a-c operated power supplies may be divided into four main parts or elements, exclusive of the regulating circuits. These are the transformer, the rectifier, the filter, and the voltage-divider system. Each of these may not be exactly alike in all systems, but regardless of their individual design, they appear in each unit in one form or another. Symbolically shown, they are as in Fig. 10-121, A, B, C, and D.

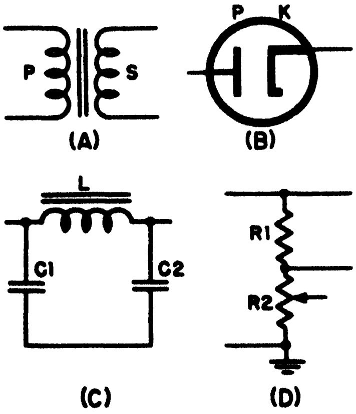


Fig. 10-121.—Typical power-supply components.

The Transformer

The power transformer (A) furnishes the voltage for the rectifier tube heaters and the other filaments as well as one or more voltages which are to be rectified. The latter are known as "high" voltages, although the meaning of "high" in these instances only distinguishes this winding from the other windings which are part of the transformer. Such transformers may contain a multiplicity of windings so as to serve a number of rectifiers of various kinds. They may contain as many as a half-dozen windings for the filaments of the rectifier tubes, the heater of the cathode-ray tube, as well as for the heaters of the tubes used in the remainder of the oscilloscope, in addition to several windings with different voltage ratings for the different rectifiers. Sometimes a separate transformer is used to furnish the very high voltage for the post-acceleration voltage rectifier.

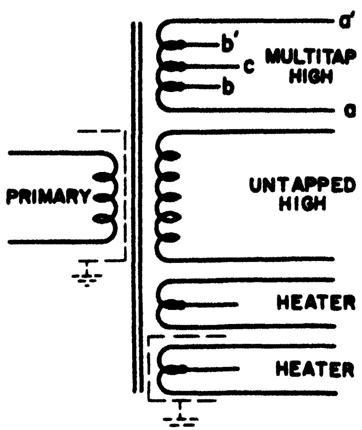


Fig. 10-122.—Typical power transformer.

A typical transformer of this kind is shown in Fig. 10-122. This particular transformer contains a single primary winding and four secondary windings. Two of these are conventional center-tapped windings for heater voltage and current supply. One of the windings affords two values of high voltage for two different kinds of rectifiers. Hence the common center-tap *c* and the two pairs of taps *a-a'* and *b-b'*. Although not a must, center-tapped high-voltage windings usually are used with full-wave rectifiers. The remaining untapped winding is a high-voltage winding intended for use with a half-wave rectifier.

Understandably, the variety of power transformers are myriad and little is gained by showing various combinations of windings. A few examples will appear later as part of the discussion of the different arrangements of half-wave and full-wave rectifier combinations.

Much may be written about the design of these transformers for use in cathode-ray oscilloscopes but we shall restrict these comments to the minimum since the design of components is beyond the province of this text. Nevertheless a few pertinent facts may be of interest. The different windings bear voltage and current ratings and are used accordingly. This means that an excessive current drain, although at the correct voltage, will reduce the voltage output from the winding because it increases the voltage drop across the resistance of that winding and because it overloads the transformer. On the other hand, a current drain less than the rated value will afford an output voltage slightly higher than the rating. Since each of the windings supplies voltage and current to a particular part of the oscilloscope, the circuit requirements are arranged to conform to the transformer ratings. The high-voltage windings also bear current and voltage ratings which are related to the current and voltage ratings of the associated power-supply output. As a rule the high-voltage winding ratings exceed the power-supply output ratings by a substantial amount so as to accommodate the different voltage drops across resistive elements, such as the rectifier tubes, the chokes, and the voltage-dropping resistors inherent to every power supply, and still furnish the rated output.

Among the features of transformer design, pertinent to successful oscilloscope use is electrostatic shielding between the primary winding and the remainder of the transformer, and also similar shielding of the winding which serves the heater of the cathode-ray tube. Both of these shields are grounded and their purpose is to eliminate capacitive coupling between these windings and the other windings on the transformer. This prevents the coupling of undesired signals from one circuit to the other. Some systems may involve pulses as, for example, the sweep circuits, and these variations in voltage may, if coupled into other circuits, modulate their currents and materially influence their behavior. This is especially true of the cathode-ray-tube heater circuit. The presence of undesired signals in the heater system can modulate the beam and give rise to false indications on the screen of the tube.

The Rectifier Tubes

Two fundamental varieties of rectifier tubes are found in oscilloscope power supplies. These are the high-vacuum and the gaseous or mercury vapor types. The high-vacuum tubes are available as half-wave and full-wave rectifiers whereas the mercury vapor tube is to be used only as a half-wave voltage converter. When symbolized, these tubes appear very much alike, since the symbol for the plate and cathode is the same in all types of tubes. Usually the presence of gas in the envelope is indicated by a dot within the circle which symbolizes the envelope.

Whichever type of tube may be used, each bears an input voltage and output current rating. These stipulate the continuous service status, that is, the maximum rms voltage which may be applied between plate and electron emitter and the maximum rms current which may be passed through the rectifier during continuous operation. In addition to the continuous duty rating, maximum peak voltage and current ratings are also given.

Since the rectifier tubes secure their voltage for rectification from the power transformer, the current and voltage ratings of the high-voltage windings of the power transformer are related to the rectifiers used with them. As a rule the current and voltage ratings of the power transformer exceed the power-supply output ratings in order to allow for voltage drops at the various resistive sections of the power-supply system.

Relative to the listing of the rectifier tubes, the following is a resume of the varieties available under the two main groups.

Gaseous	*High Vacuum*
Half wave	Half wave
High voltage	Low voltage
	High voltage
	Full wave
	Low voltage

Concerning the above listing, the meanings of high voltage and low voltage are quite broad. They are

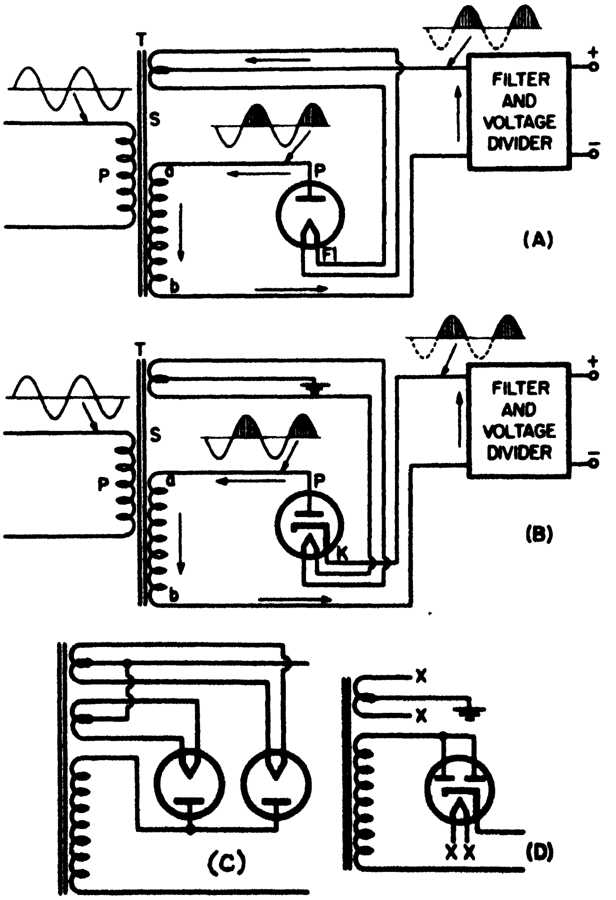


Fig. 10-123.—Half-wave rectifier action, (A) direct-heated cathode, (B) indirect-heated cathode, (C) and (D) are variations of (A) and (B).

strictly relative terms, for a rectifier rated at 1,000 volts will be high voltage in comparison to another which is rated at 300 volts and designated as low voltage, and in turn, will itself be low voltage when compared with a rectifier which will accommodate 5,000 to 10,000 volts. Practically speaking, the rectifiers designated as low voltage and found in oscilloscopes operate at voltages up to 300 or 400 volts; those which rectify voltages generally in excess of this value and up to several thousand volts are spoken of as high voltage.

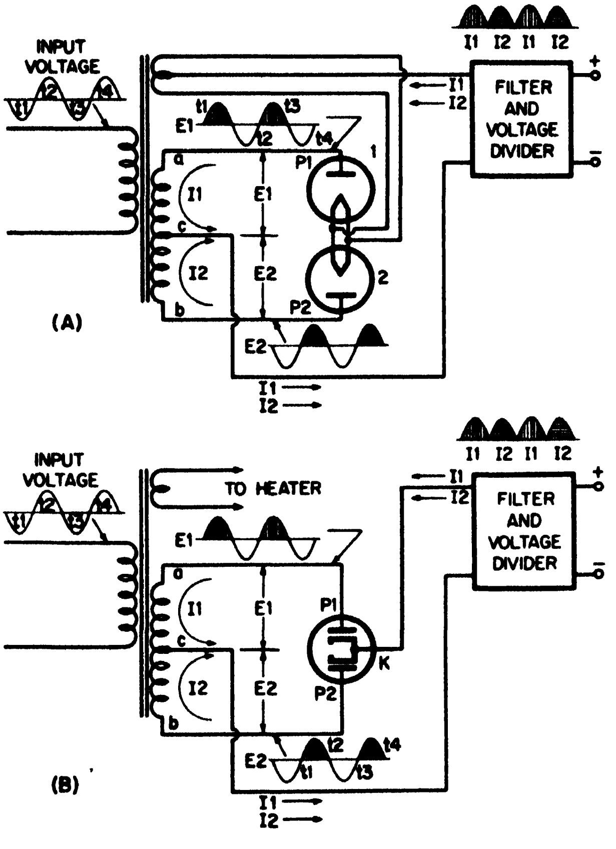


Fig. 10-124.—Full-wave rectifier action, (A) direct-heated cathodes and (B) indirect-heated cathodes.

When symbolized, the half-wave rectifier is shown as consisting of a plate and a cathode or a filament as in Fig. 10-123A, whereas the full-wave rectifier contains two plates and a common cathode or a single filament which is common to both plates. This is shown in Fig. 10-124B. Obviously the full-wave rectifier tube is nothing more than two half-wave sections contained in a single envelope with such physical changes in the electron emitter as will satisfy the current needs of the two plates. This is a commonplace form for the high-vacuum type tube but is not found in the mercury vapor kind; the latter variety always comes as a half-wave rectifying element.

Sometimes two individual half-wave rectifiers are connected in such fashion as to perform the functions of a full-wave rectifier, and conversely, a full-wave rectifier tube may have its plates connected in parallel so as to perform as a half-wave rectifier. Then again two individual half-wave rectifiers may be connected in parallel, that is, the two plates are joined externally, as are the two cathodes, or the two filaments are placed in parallel. When two individual tubes are used in this fashion, the current output is increased but the voltage is not. Moreover the tubes must be similar, with like current and voltage ratings.

An increase in current output, as well as in voltage, is obtainable by the use of a full-wave rather than a half-wave rectifier, as will be shown later, but quite frequently the voltage and current demands are met by the half-wave system, which is more economical. Therefore the half-wave rectifier finds very frequent use in oscilloscopes, especially when the current drain is comparatively low.

Although not evident in the drawings, the cathode type tubes are equipped with heaters just as in the conventional type of receiving vacuum tubes. As to the manner of performance of all of these tubes, one electrical condition is paramount. Since each of these tubes is the equivalent of a diode, conduction (plate current flow) takes place only when the plate or anode is made positive relative to the electron emitter, be it a cathode or a filament. After the polarities are reversed, that is, when the cathode is positive with respect to the plate, conduction ceases and the tube becomes a nonconductor. Thus, rectifier tubes are unilateral conductors of current, although it must be understood that, when conducting, each type of tube presents a certain amount of internal resistance, and therefore, a certain amount of voltage drop. In the case of the high-vacuum tube, this is variable over a relatively limited range determined by the magnitude of current. In the case of the mercury vapor rectifier, the voltage drop is substantially constant at about 15 volts over the complete range of output current available with the tube.

With this much behind us, we can examine the two fundamental varieties of rectifying systems. These are the half-wave and full-wave circuits. Inasmuch as the rectifying system is not influenced by the filter arrangements, rather the reverse is true, we can hold the examination of the filter and the voltage divider in abeyance until later. When dealing with the rectifiers, we will show these two parts of the power supply as a single block.

The Half-Wave Rectifier

Two examples of the elementary half-wave rectifier are shown in Fig. 10-123. They differ only in the kind of emitter used; one employs a filament whereas the other utilizes a cathode for that purpose. In the former, the filament is a part of the rectifying system, while in the latter instance, the heater similarly represented serves merely as a means of raising the cathode to an electron-emitting temperature.

Concerning the power transformers, the ones shown are typical of some installations whereas in others the secondaries may be just two of many wound on the same core.

As to the process of rectification, 2 cycles of the input line voltage are shown above the respective primaries. In like manner 2 cycles of the induced "high" voltage are shown above each of the two related secondary windings. The 180° reversal of phase between the primary and secondary voltages is natural in a transformer. The low voltages applied across the filament and heater, respectively, require no attention.

The polarity signs which accompany the secondary voltage waveforms indicate the relative instantaneous polarities of the points *a* and *b* on the two secondaries. Point *b* in (A) can be assumed joined to the filament of that rectifier tube, and point *b* in (B) can be assumed joined to the cathode. Actually these connections are not so direct, but for purposes of explanation, it may be assumed so.

Analyzing diagram (A), the first half-cycle of the input voltage results in a secondary voltage which makes the plate *P* negative relative to its filament. Consequently, conduction does not take place in the rectifier tube because the negatively charged plate does not attract electrons from the space charge within the tube. Therefore, the system on a whole is inactive during the negative alternation of the secondary voltage, or the positive alternation of the primary voltage. Since all of our interest lies in the secondary circuit of the transformer and the rectifier connected to it, and knowing the relationship between the primary and secondary voltages, we can temporarily neglect the action in the primary circuit.

During the next alternation or half-cycle of the secondary voltage, which we show by vertical shading lines, the plate of the rectifier is made positive with respect to its filament, and a pulse of current passes through the system as indicated by the arrows. Neglecting for the moment the action of the filter system, we can show this pulse of current to have a shape somewhat like the voltage alternation which was rectified. Actually it is somewhat different in shape but this is irrelevent to the point we desire to convey. The output current pulse is shown in vertical shading lines immediately above the filter in (A), and the dotted alternation represents the time interval corresponding to the negative half-cycle of the secondary voltage.

The next instant the plate again is negative relative to its filament and conduction ceases and remains so during the entire negative alternation of the voltage applied to the rectifier. Then polarity of the secondary voltage reverses and again the plate of the rectifier is positive and another pulse of current passes through the system as indicated. These alternate periods of conduction and nonconduction continue as long as the voltage is applied from the secondary winding and the direction of each current pulse is the same as the previous one. As can be seen, they occur once during each cycle of applied voltage, always during the positive alternations, so that only one-half of the applied voltage wave is being rectified. Hence the name of half-wave rectifier.

Concerning the direction of movement of the charges through the system, the representation of the output current pulses shows them to be always in the same direction and always above the zero reference axis or in the positive zone of the graph. Thus the process of rectification might be described as being an arrangement whereby the positive alternations of a voltage are permitted to pass through a system and the negative half-cycles are cut off. Later on you will see that the positive pulses of current are used to charge a capacitor.

If we correlate the output current pulses with the voltage input to the rectifier, we note that both are cyclic in character; that the number of output current pulses which occur each second in a half-wave rectifier are the same as the frequency of the voltage output to the rectifier, or as the frequency of the primary voltage. This is a fundamental condition. In addition to this, the output current pulse is a distorted version of the input voltage, and contains a series of harmonics. All of these combined constitute the "hum" voltage. More about this later.

Concerning the half-wave rectifier in (B) of Fig. 10-123, its basis of operation is exactly the same as outlined for arrangement (A) except for the use of a cathode *K* in place of the filament *F*. In all other respects, the two operate in identical fashion.

The use of two half-wave tubes in parallel, or the two plates of a full-wave rectifier in parallel, as shown in (C) and (D), respectively, of Fig. 10-123 can be seen to introduce no major variations in the fundamental principles of operation. The omission of the heater winding wiring in (D) is simply for convenience in drawing.

The Full-Wave Rectifier

The full-wave rectifier is slightly more elaborate than the half-wave system. It involves a change in the transformer winding which feeds the rectifier tubes, and also the use of two rectifying sections. As we stated earlier, these may be individual tubes or they may be contained within a single envelope. Schematically the two arrangements are shown in Fig. 10-124A and B, where they are organized for comparison with the symbolized schematics of Fig. 10-123.

The full-wave rectifier consists of two half-wave rectifiers which function on opposite alternations of the applied voltage. Let us analyze this action in the circuit (A) in Fig. 10-124. This is directly comparable with circuit (A) in Fig. 10-123.

The rectifier tubes are two individual half-wave converters and they secure their filament voltage from a common transformer winding. The high-voltage secondary is a continuous winding which is center-tapped so that two equal voltages *E1* and *E2* are available from the output. Since each of these voltages is intended to function with an associated rectifier tube, the value of the voltage equals that which normally would be obtained from an untapped winding used with a half-wave rectifier tube. The center-tap connection passes into the filter where, in effect, it joins the center tap of the filament winding.

The tap at the electrical midpoint of the high-voltage winding furnishes two voltages which, at any one instant, are 180° out-of-phase relative to the center tap. This is indicated by the waveform curves marked *E1* and *E2*. The polarity of one end of the complete secondary winding is opposite to that existing at the same instant at the other end. Therefore, if we assume that at one instant point *a* is positive, everything else along the winding is negative, which, of course, includes the center tap *c*. On the other hand, the next instant, when the polarity of the output voltage is reversed so that point *b* is positive, every other point along the winding is negative, and again this includes the center tap *c*.

Accordingly points *a*, *b*, and *c* possess dual polarity, but at any one instant, these polarities are either positive or negative relative to each other. If we select the center tap as the reference point, it retains its negative polarity with respect to *a* when that point is positive and with respect to *b*, when that point is positive. Therefore, we can consider the output available from the high-voltage winding as consisting of two individual voltages which alternately become positive and negative at the ends which are connected to the two plates *P1* and *P2*.

During the negative alternation *t1* of the primary voltage, the first alternation of the secondary voltage *E1* is positive. The plate *P1* is then positive relative to its filament and the current pulse *I1* flows through the system, as shown by the arrows which bear that label, and by the output current pulse waveform which is similarly marked. At the same time, the first alternation of voltage *E2* is negative relative to the common reference point, so that the plate *P2* connected to point *b* is negative relative to its filament. Therefore, conduction does not take place between *P2* and its filament but does between *P1* and its filament.

During the second alternation *t2* of the first cycle of primary voltage, the polarity of the secondary voltage reverses, making point *a*, and therefore, plate *P1*, negative relative to its filament, and point *b* and plate *P2* positive relative to its filament. Conduction now takes place in tube *2* but not in tube *1*, and a second current pulse *I2* flows through the system in exactly the same direction as the previous current pulse. The use of similar shading of the related alternations of secondary voltage and the output current pulses aids recognition of the related actions.

It is evident from what has happened during the two alternations of the first cycle of primary voltage that conduction takes place alternately in the two rectifiers, so that both alternations are rectified. Hence the name of full-wave rectifier. The action during the time intervals *t3* and *t4* of the input voltage cycle are repetitions of what we described in connection with the two alternations periods *t1* and *t2*.

Comparing the output current pulses for the half-wave and the full-wave rectifiers discloses that, in the latter, a current pulse exists for each alternation of the input voltage, and if we view these current pulses as being cyclical in character, their frequency is twice the frequency of the primary voltage. If the line voltage or primary voltage frequency is 60 cps, the cyclic variation of the output current pulses in the half-wave rectifier occurs at the same rate, whereas in the full-wave rectifier, it occurs at the rate of 120 per second.

Still another point of interest becomes evident by this comparison. The full-wave rectifier affords two output current pulses, whereas the half-wave system affords only one in the corresponding period of the input voltage. Therefore, with everything being equal, the output current in the full-wave system is twice that of the half-wave arrangement.

Concerning the modified rectifier tube arrangement shown in (B) of Fig. 10-124, its rectifying action is identical to that of system (A). The use of a cathode-type tube which contains both plates in the same envelope introduces no modifications. Each plate acts independently of the other in conjunction with the common cathode *K* in exactly the same manner each of the plates in the individual tubes of arrangement (A) acted in concert with its individual filament. Since these filaments were electrically common relative to the path of the charges in the output current pulses, they are the equivalent of the single common cathode *K*. The behavior of the heater winding in (B) may be ignored since the source of these charges is the cathode; the heater simply raises the temperature of the cathode to the electron-emitting level.

Filter Systems

Filter systems used in power supplies may be of very many kinds, but three varieties are fundamental. These are shown in Fig. 10-125. These may be combined in many ways, but any such combination can be analyzed into these three basic forms. Systems (A) and (B) are known as the π (pi) type, whereas (C) is representative of the *L* type. As shown, each comprises a single section, but every complete filter need not consist of one section. Several may be used so as to most effectively accomplish their purpose.

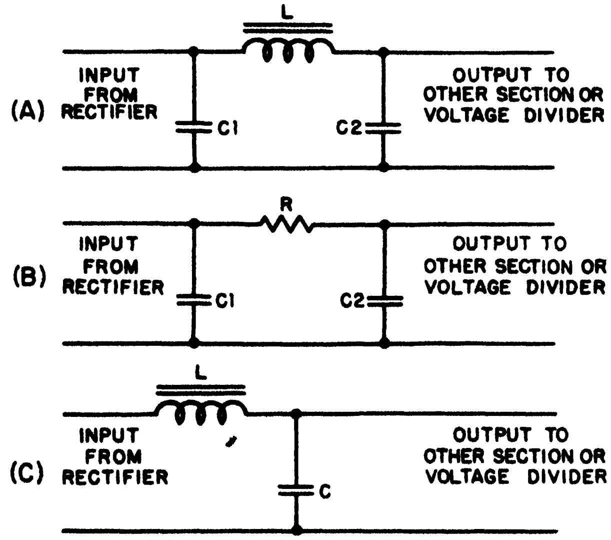

L
INPUT FROM RECTIFIER
(A)
C1
C2
OUTPUT TO OTHER SECTION OR VOLTAGE DIVIDER
R
INPUT FROM RECTIFIER
(B)
C1
C2
OUTPUT TO OTHER SECTION OR VOLTAGE DIVIDER
L
INPUT FROM RECTIFIER
(C)
C
OUTPUT TO OTHER SECTION OR VOLTAGE DIVIDER

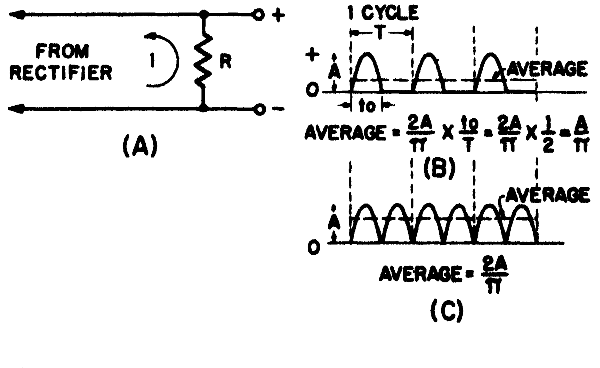
Fig. 10-125.—Fundamental filter sections, (A) π type using an inductance, (B) π type using a resistance, and (C) *L* type.

Concerning the organization of the individual sections of power-supply filters, one specific condition is adhered to at all times. This is the use of conducting elements *L* or *R* in the series arms and the use of *C* in the shunt arms. The reason for the former is self-evident; these units are the paths for the direct current secured from the rectifier system, and therefore must permit the passage of such current. This does not mean that only *L* and *R* will be found associated with the series arms of such filters. Sometimes *C* is also used but only as a means of resonating *L* so as to make the filtering action more effective. When so used, it still is not in the path of the direct current; the element *L* forms that path. The element *C*, on the other hand, usually is the shunt arm.

The three fundamental filter arrangements shown in Fig. 10-125 can be classified under two circuit groups. These are *capacitance input* (A) and (B) and *inductance input* (C). The identification is based upon the electrical element that faces the rectifying tubes, or, we might say, the element most closely related to the behavior of the conduction current pulse. These are *C1* in circuits (A) and (B), and *L* in circuit (C). In the capacitance-input circuits, the current pulse charges the capacitor, and in the inductance input, the inductor *L* controls the amplitude of the current pulse variations which charge the capacitor that follows the inductor. The behavior of the input capacitor in the capacitance-input type of filter is very interesting.

Capacitance-Input Filter Action

The action of the filter system can be realized by reference to Fig. 10-126. The power supply is a storehouse of electrical energy. Its function is to be a source of constant d-c voltage at a specified current drain. In order to show how all of this is accomplished, examine arrangement (A). Here a resistor *R* takes the place of the complete filter block shown in Fig. 10-123. Because of the direction of the motion of the charges through the element, the polarity indicated is developed. This


FROM RECTIFIER
I
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(A)
1 CYCLE
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+
A
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AVERAGE
t_0
AVERAGE = $\frac{2A}{\pi} \times \frac{t_0}{T} = \frac{2A}{\pi} \times \frac{1}{2} = \frac{A}{\pi}$
(B)
AVERAGE
A
0
AVERAGE = $\frac{2A}{\pi}$
(C)

Fig. 10-126.—(A) Load on an unfiltered rectifier, (B) voltage across load for half-wave output, and (C) voltage across load for full-wave output.

current occurs in the form of pulses, one pulse for each cycle of input voltage in the half-wave rectifier and two such pulses during each cycle of input voltage in the full-wave rectifier. These are shown as (B) and (C), respectively.

Now it is evident that the voltage pulses which would develop across the terminals of *R* when the current pulses are of the form shown would be anything but constant. Although these current pulses have an average value, and the resultant voltage may be identified in terms of this average value, the fluctuations in the output voltage would defeat the use of the rectifier system as a power supply. The peaks of the pulses must be flattened and the hollow parts filled in so that the fluctuations are removed, and a constant output voltage is thus made available. This is done in the filter.

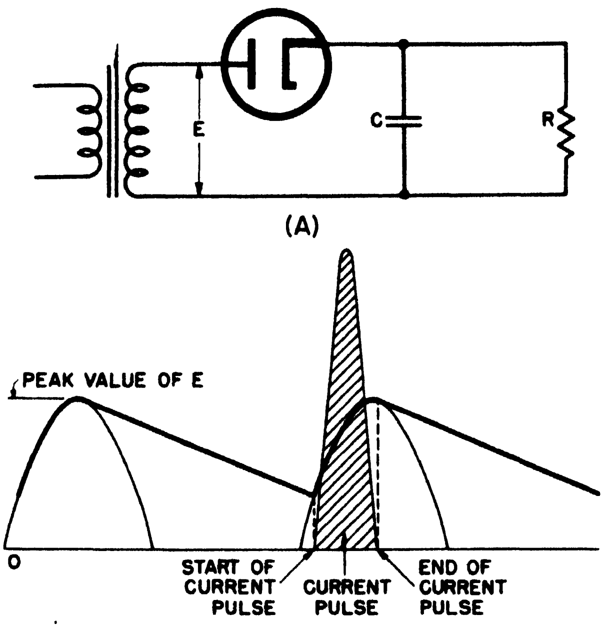


Fig. 10-127.—(A) Simple capacitance-input filter with a half-wave rectifier, and (B) output voltage waveform (heavy), and occurrence of the current pulse.

In Fig. 10-127A is shown a capacitor *C* placed across the load resistor. This corresponds to the input filter capacitor of Fig. 10-125. Since the ohmic value of *R* is appreciably greater than the reactance of *C*, the current pulses from the rectifier charge the capacitor to a voltage which approximates the peak value of *E*, being less than that amount by the voltage drop across the internal resistance of the rectifier tube. During the negative alternation of *E*, the capacitor cannot discharge through the vacuum-tube circuit because the cathode is positive relative to the plate and conduction is not possible. It, therefore, discharges through *R*, as shown by the solid line in (B). As charge is drawn from the capacitor, its voltage naturally falls. The extent of this decrease in voltage is determined by the magnitude of the current drain through *R* or whatever is the load, and also by the value of *C*; for any given current drain, the greater the value of *C*, the smaller is the decrease in voltage across *C*. The smaller the current drain between charging periods, the smaller is the decrease in voltage.

Assuming one set of conditions, the capacitor *C* continues to discharge until the next positive current pulse comes along, at which time the capacitor once more is charged. Now it is interesting to note that in (B) the voltage across the capacitor has not fallen to zero by the time *E* again makes the plate of the rectifier positive relative to the cathode. This means that the current pulse cannot start flowing until the voltage *E* exceeds the voltage on the capacitor. The smaller the decrease in voltage across the capacitor, the later is the starting time of the current pulse during the positive alternation of *E*.

Once the current pulse starts to flow, the charging process also starts, and, in time, again charges the capacitor to its proper value. But shortly after the peak value has been reached, the fall in voltage *E* is more rapid than the rate of discharge, so that again the voltage across the capacitor exceeds the value *E* and the conduction ceases. This cannot help but lead to the conclusion that after the first few cycles of operation have been completed, the current pulses which constitute the charge current occur as sharp pulses. Moreover, the smaller the discharge of the capacitor the shorter in duration these pulses of charging current.

Now if we examine the voltage curve in (B) (heavy line), it is evident that the capacitor is behaving like a storehouse of energy; but the voltage output is still not constant, it still fluctuates. It is also evident that if the gap between periods of charging could be filled in, the fluctuation in output voltage would not be as great. Of course, if the value of *C* were increased very greatly and the discharge rate were kept the same, the slope of the voltage curve would not be as great. There is, however, a limit to the value which may be used for *C* because of a related effect. This is the magnitude of the current pulse at the instant the charging process starts. Rectifier tubes bear peak-current ratings and these may be exceeded if the value of *C* is too large. Thus it is a matter of finding a compromise between the capacitance of *C*, the permissible peak value of charging current, and the current drain. In fact the very presence of a large charging current when a capacitor is the immediate load on a rectifying system is what prevents

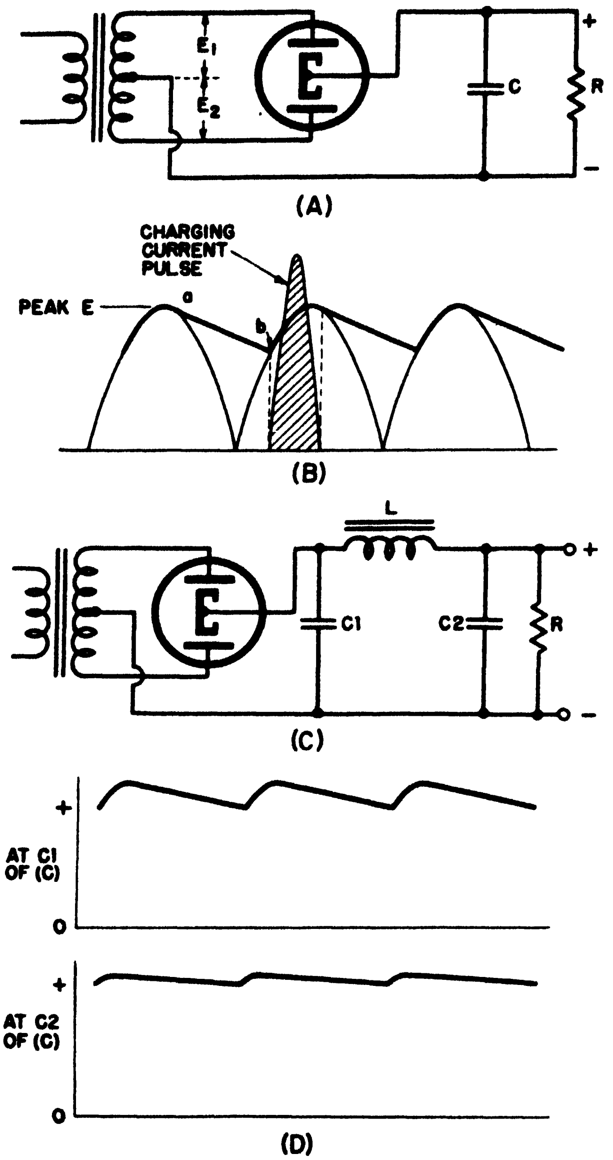


Fig. 10-128.—(A) Simple capacitance-input filter with a full-wave rectifier, (B) output voltage waveform (heavy), and occurrence of the current pulse, (C) full-wave rectifier with π type filter, and (D) voltage waveforms at input and output, respectively, of filter in (C).

the use of capacitors as the first element in the filter when a mercury vapor type of rectifier is used. Excessive charging current through the tube may irreparably damage it.

The behavior of the capacitor when used with a full-wave rectifier as shown in Fig. 10-128 is fundamentally the same, except that now, as illustrated in (B), the time for the voltage across the capacitor to fall has been reduced to half of what it was in the half-wave system. This means: 1. a higher voltage is available from the filter system; 2. a higher rate of variation of the output voltage; and 3. a reduced amount of fluctuation in voltage for any one value of current drain. It must be understood, of course, that the above reference to a higher voltage across capacitor *C* does not mean that a change has been made in the value of the a-c voltage to the rectifier. *E* in Fig. 10-127 has the same values as *E1* and *E2* in Fig. 10-128A. It is simply that the average voltage across *C* between the peak value *a* and the final value *b*, when the next charging period comes along, is greater in the full-wave rectifier than in the half-wave rectifier. This is easily seen by comparing these two points along the voltage curves in the aforementioned two illustrations.

The pulsating voltage across the capacitor *C*, which can be viewed as the form of the voltage across *R* is a form of distortion. While these voltage fluctuations have a fundamental as we described before, they also contain numerous other frequency components. All of these combined constitute the *ripple* in the output, or, as some might call it, the *hum* in the output. Anything that is done to remove the voltage fluctuations and so furnish a constant d-c output will automatically reduce the amplitude of these ripple components and remove the hum present in the supply. This is a function of the filter, but since the capacitor *C* in the two illustrations just discussed also is a part of the filter, the dual function of the latter is evident.

The remainder of the elements of the filter with capacitance input are responsible for the removal of the voltage variations across the input capacitance as the power supply furnishes voltage and current to a load. As a basis for discussion, let us add the remaining elements of a single-section filter to the rectifying system, as shown in Fig. 10-128C. The current and voltage conditions shown at (B) are assumed to exist across *C1*, the input capacitor. The path through which the input capacitor *C1* will discharge now contains the inductor *L*. We know from basic fundamentals that the inductance of the winding will tend to prevent rapid changes in current; it will tend to prevent the rise of a current, and if one exists, it will tend to keep that current flowing, if for some reason it is caused to stop. The iron core in inductor *L* is intended to make this inductive effect greater, as great as possible consistent with the other requirements of design. Now it is to be noted in circuit (C) in Fig. 10-128 that another capacitor *C2* also is a part of the filter network; that, in effect, it is in shunt with capacitor *C1*, with *L* as a part of the circuit which links the two capacitors.

In view of the circuit arrangement, the input capacitor can be charged by the current pulse, but if capacitor *C2* is to be charged, the charging current must first.

pass through *L*, which, in turn, tends to hold down any sudden changes in this current. In other words, the rapid fall in voltage across *C1* is not transferred to *C2* because the variations in the correspondingly changing current will not be transmitted. The inductor *L*, therefore, tends to supply to capacitor *C2* a charging current which is more constant than that flowing out of *C1*. The final result is that capacitor *C2* also behaves like a reservoir of electrical energy and the fluctuations in voltage across its terminals, with current drain on the power supply, are very substantially reduced. This may be illustrated as in (D) of Fig. 10-128.

By using another section, these small fluctuations are further reduced. The system can be made so that the ripple (fluctuations) voltage does not exceed, say, a very small fraction of a percentage of the rated output of the power supply. Whether a second filter section is required depends a great deal upon the nature of the load. Some amplifier systems are such that a fluctuation in operating plate voltage has opposite effects on the two tubes in a stage, and tend to cancel each other. In still other applications, a small variation in operating plate voltage is tolerated because it is not harmful.

In still other instances where a high order of constancy is needed, especially where the current drain may vary over a wide range, voltage stabilizers or regulators are used. This will be discussed in brief later in this chapter.

Inductance-Input Filter Action

The behavior of the inductance-input type of filter can be understood by reference to Fig. 10-129. As we stated before, the inductive effect in a winding tends to resist sudden changes in current and when the driving voltage no longer exists, the inductance tends to maintain the decaying current. Therefore, if a choke *L* is placed in series with a resistance *R* across a rectifier system as shown in arrangement (A) of Fig. 10-129, the current pulses are modified in form from the light line, which represents the current when the load is purely resistive, to the heavy line. As can be seen, abrupt changes in the current pulses have been prevented. As a matter of fact, the action of the inductance tends to prevent the rise of the current in one direction and the fall of the current in the other.

This tendency on the part of the inductance creates a number of related conditions which are very important. The first is that it tends to maintain current flow in the rectifying system at all times and, in addition, prevents the rise of the current to that peak value which would prevail if a capacitor instead of an inductor were the element at the input of the filter. In other words, the abrupt rise of charging current with capacitive input does not take place with inductance input and it is this condition which demands the use of inductance input with mercury vapor rectifiers. These tubes cannot tolerate sudden surges of current.

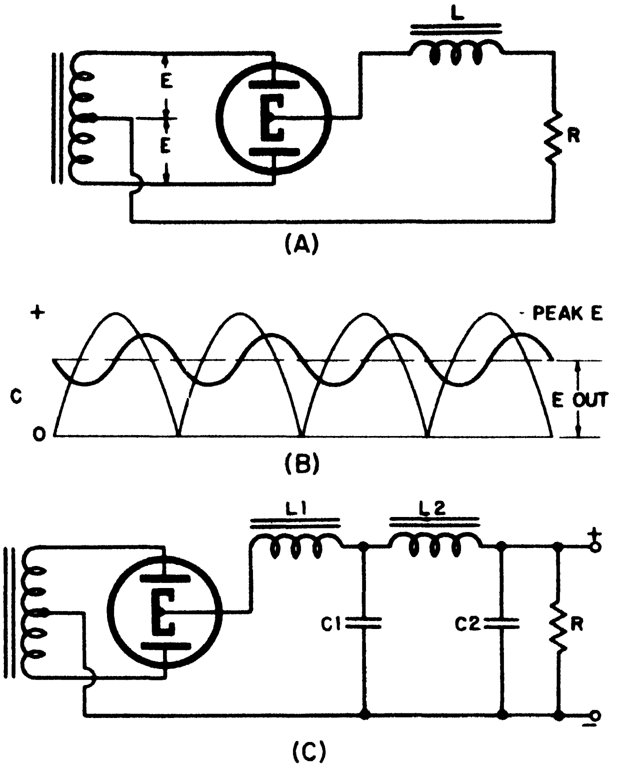


Fig. 10-129.—(A) Simple inductance-input filter with full-wave rectifier, (B) output voltage waveform of (A), and (C) double *L* type filter system with full-wave rectifier.

Now, consistent with the absence of these peak currents, it naturally follows that the rectifier output voltage cannot reach the peak value of the a-c voltage input, as in the capacitance-input filter. The final result is that the output voltage available from a power supply with an inductance-input filter is substantially less than that from a capacitance-input filter supply with like values of a-c voltage input to the rectifier and like current drain on the supply.

The simple use of an inductance in place of a capacitor at the input of the filter does not necessarily mean complete filtering, even though the former system affords a more constant voltage across the input of the system. Capacitance is still required for the storage and release of the energy. This is the purpose of the capacitors in the filter system as shown in (C) of Fig. 10-129. The equivalent of the input capacitor in the capacitance input filter is *C1* in Fig. 10-129 (C), except that the conditions of charge and discharge take place with less violent fluctuations, since the inductor *L1* tends to keep the current during the charging period more constant, and inductor *L2* tends to keep the discharge current more constant. Capacitor *C1* tends to maintain its charge more uniformly as does capacitor *C2*. The sum and substance of such an arrangement is that the inductance-input type of filter affords better *regulation* in the power supply. By this is meant the relative change in output voltage with different current drains. It is expressed in percentage of the rated output voltage. Substantial changes in current drain applied slowly or abruptly do not cause major changes in the voltage output even without the use of gaseous or high-vacuum voltage regulator tubes. The voltage output will of course change, but to a very much smaller extent than in capacitance-input filter power supplies which also do not contain voltage regulators.

Resistance-Capacitance Filter Action

The use of resistance in place of inductance in the capacitance-input type of filter as shown in (B) of Fig. 10-125 and in Fig. 10-130 is an economical way of accomplishing a power-supply filter. It affords filtering action, but not as effectively as its *L-C* counterpart.

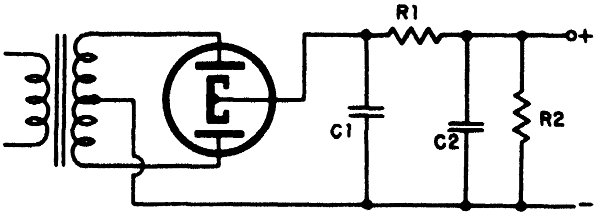


Fig. 10-130.—Full-wave rectifier with π type filter using a resistance.

To appreciate its action, consider the circuit of Fig. 10-130. The behavior of *C1* is as described in the *L-C* filter. By virtue of its energy-storing action, *C1* tends to provide a certain amount of filtering or "smoothing." This could be improved by making the value of capacitance greater, for then the loss in voltage with unit current drain would be less than with a small value of *C*. But an improvement in performance is accomplished if the added *C* (*C2*) has an appreciable amount of resistance in series with it, as shown in the diagram. This stems from the relative reactances of *C1*, *C2*, and *R2* to the frequency of voltage fluctuation. For example, for any one value of voltage across *C1*, it may be viewed as being made up of a d-c component and an a-c component. The series resistance *R1* drops both of these voltages somewhat and the division of the two components relative to the two parallel paths, *R2* and *C2* is a matter of the relative reactance and resistance.

By making the reactance of *C2* very small, effectively a short circuit at the fundamental ripple frequency, but effectively an infinite impedance to d.c., and since the resistance of the load *R2* is finite for both, the presence of *C2* very greatly reduces the ripple in the voltage so that it behaves as a filter. The filtering action so obtained is not the best possible, although the use of two such sections when the d-c voltage drop across the series arms can be tolerated, is capable of a very substantial reduction in ripple. Sometimes economy in the cost of the filter is accomplished by the combined use of an *L-C* section and an *R-C* section, the former being the first section nearest the rectifying tubes.

The presence of resistance in the series branch of the filter dictates certain conditions of operation, essentially that the current drain on the power supply be low. In view of the need for a substantial amount of resistance for this element, 40,000 to perhaps as high as 100,000 ohms is in order to make *C2* effective; the current drain through this resistor must be small, otherwise a very large voltage drop will occur across this element. If it can be tolerated, all well and good, but in most instances, high voltage as well as the required current drain are requirements of power supplies, so that any such loss in voltage is sheer waste.

The second point of importance is that while operable, such a filter will not afford an output voltage free from ripple. This means that the filter can be used only in those power supplies which feed devices that can accommodate a substantial amount of hum. Finally these systems are useful mostly in circuits where the current drain is substantially constant and for which the filter system has been designed. Substantial increase in current drain will adversely affect the general behavior of the filter and the power supply.

The Voltage Divider

The voltage-divider part of the power supply performs a function which is described by its very name. Accordingly, it is supposed to divide or allocate the available voltage in accordance with the needs of the system connected across it. While we have shown numerous examples of a resistor *R* as the termination of the filter and have referred to it as the "load" on the power supply, in reality, the load on the power supply is the equipment which the device supplies with voltage and which absorbs the current drain.

As to the *division* of voltage available from the power supply, sometimes that is done, and sometimes the full voltage is applied to one point. This results in the use of two kinds of so-called divider systems. Actually only one of them is a divider. What we mean is shown in Fig. 10-131A and B.

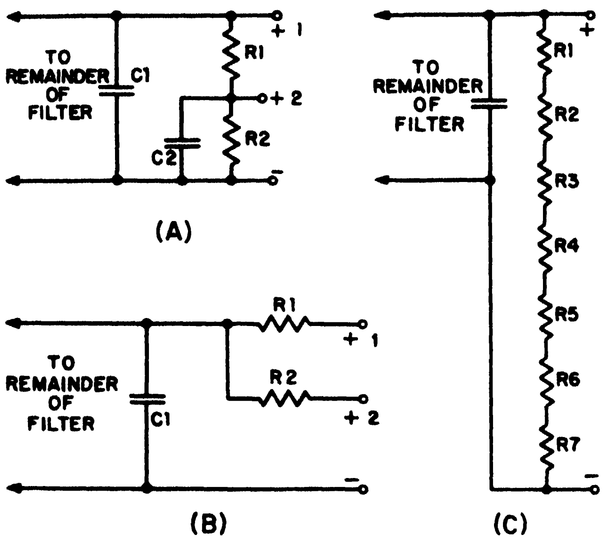


Fig. 10-131.—Various kinds of voltage dividers.

In the former (A), the divider network consists of resistor *R1* and *R2* in series connected across the output of the power-supply filter. By virtue of the direction of current flow, the polarities shown are established. The voltage impressed across the series network is available between the minus terminal and the uppermost positive terminal. The lower positive terminal affords a voltage which is less than the maximum by the drop across *R1*. This much is basic. Now the interesting detail about such a voltage divider is that it is known as a "bleeder" as well, for it does act to "bleed" off the charge remaining on the filter capacitors in the event that the rectifier is turned off simultaneously with the devices which draw current from the power supply.

At the same time, such a system also is a fixed load or fixed current drain on the supply, for no matter what other equipment may be connected to the power supply, this pair of resistors will continually draw current from the filter system as long as the rectifier is operating. By connecting the equipment leads to points *1* and *2,* the voltage available at these points is applicable to the equipment.

When viewed as a voltage divider, the sections of such a system will divide the available voltage in proportion to the relative resistance values provided that only one load exists across the divider, namely, the main load across the entire divider. Regardless of what this current drain may be, assuming, of course, that it is within the rating of the device, the voltage at the intermediate terminals on the divider (which may be more numerous than we show in the illustration) will be divided in proportion to the resistance elements in the series. But, if a load is applied between the negative terminal and any one of the intermediate positive terminals, the division of voltage at the intermediate terminals no longer will be in proportion to the resistances in the series because of the unequal currents passing through the series elements. Only when the main load is connected, the current is the same throughout the divider network, which no longer is true if the intermediate terminals also are in use.

Another type of voltage distribution arrangement is shown as (B) in Fig. 10-131. Here the resistors *R1* and *R2* are in series with the devices which draw the current. The level of the voltage at the load point is determined by the drop across the series element due to the current passing through it. Such an arrangement is not general because it has the weakness of not bleeding the charge off the filter capacitors after the rectifier is turned off. Also, it permits a substantial rise in voltage across the filter capacitors when the load equipment is turned off.

Flash-Over in Power Supply

The high voltage required for some of the cathode-ray-tube operating potentials forces certain physical design considerations into the voltage dividers of the power supplies which furnish these voltages. We are referring to the implication of the simple series network of two resistors in (A) of Fig. 10-131. It is entirely possible that the voltage available from such a power supply might be several thousand volts, and consequently care must be exercised to see that the physical design of the power-supply bleeder or voltage divider is such as to prevent flash-over due to the high-voltage difference between the extremes of the divider. This demands a long electrical path or a long leakage resistance path. The divider of Fig. 10-131A might be represented as (C) of the same illustration. Each of the resistors *R1* through *R7* may or may not be of like value, but they may be so arranged physically as to make the path from the high-voltage point to the negative terminal the longest possible. The aggregate ohmic value can easily be attained with a single unit, but if that were done, the leakage path would be very short and flash-over a possibility.

This matter of flash-over is of more than casual interest. While it is assumed that the design of a unit is such as to adequately fulfill all of the requirements, it is conceivable that under certain conditions of operation and failure of an element, a replacement is made.

If such a replacement is made in a system where high voltages are encountered, the matter of insulation is of extreme importance. This applies to every part of the power supply, even if, at the moment, we seem to be speaking about the voltage divider. The writer's experience with oscilloscopes which employed high voltages has frequently proved the need for careful examination of insulation when remedying faults in voltage dividers. Frequently the matter of insulation on wires and "spaghetti" around power-transformer leads was a matter of concern.

As long as we are on the subject, we might mention such very simple items as insulated terminal strips which may be found in power supplies. A slight misalignment of one of the connecting terminals to which a high-voltage lead may be joined, is apt to bring that terminal so close to a grounded point as to make flashover very likely. Strangely enough, this might happen the instant the equipment is turned *off* and this makes the matter ever so confusing for it "operated perfectly on the previous occasion."

Polarity in Voltage Divider

It is commonplace practice to ground the negative terminal of the power supply, in which case, every other terminal on the voltage divider is positive. This is shown in simplified form in (A) of Fig. 10-132. If any measurements of voltage are made, they are with respect to ground, which is the negative terminal. In oscilloscope equipment, one will find the above quite frequently, but it is possible that some point other than the most negative will be at ground potential. Being at ground potential makes that point 0 voltage with respect to ground, but it may be either positive or negative relative to some other point on the divider. For example, in (B) of Fig. 10-132, ground along the divider is located at a point which is *positive* relative to the most negative point in the system, and negative relative to the positive point in the system. In other words, here is a voltage divider in which two voltages positive relative to ground and two voltages which are negative relative to ground (below ground) are available. Obviously a voltage measurement made with ground as the reference point will result in two polarities, positive at some terminals and negative at other terminals, depending entirely on the connections to points (*1*), (*2*), (*3*), or (*4*).

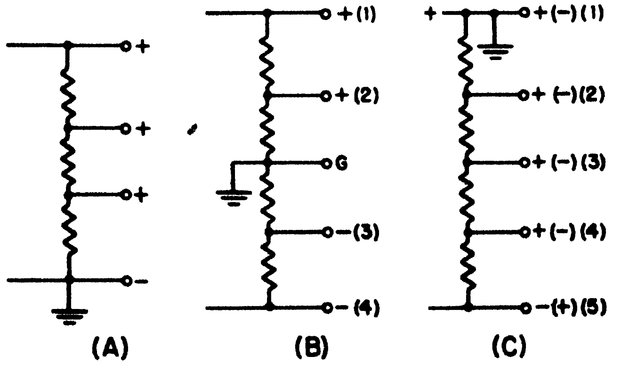


Fig. 10-132.—Illustrating polarity in voltage dividers, (A) gives a positive output with respect to ground, (B) gives both positive and negative outputs, and (C) gives negative outputs.

Connecting a circuit to the grounded terminal which we designate *G*, as a matter of convenience, makes that terminal actually positive relative to terminals (*3*) or (*4*) and negative relative to terminals (*1*) and (*2*), although all the time it is at ground potential. This is a conceivable arrangement for making the cathode of a cathode-ray tube negative relative to ground, which is the potential of the accelerating anode; the accelerating anode requires a positive potential with respect to the cathode which it receives this way, yet the accelerating anode is negative with respect to the post-deflection anode, and this allows the application of a still higher positive voltage at the post-deflection intensifier terminal. These voltages may not necessarily be obtained from a single divider in exactly this fashion, but it does indicate the possible combinations of power-supply voltages and polarities. Examples of these will be given shortly.

Still another example of varying location of the ground point on a voltage divider is shown as (C) in Fig. 10-132. Here the most positive point on the divider is connected to ground. The polarity of the remaining points is a matter of what is taken as the reference point. If ground is positive [terminal (*1*)] then all of the other points along the divider are negative, whereas if terminal (*5*) is taken as the reference point, then all of the other points along the divider are positive with respect to it. The fundamental polarity along the divider is, of course, established by the direction of the current flow through the unit. This makes terminal (*1*) the most positive and terminal (*5*) the most negative and it is not altered in the slightest by the ground connection. The change in polarity comes about through the selection of the reference point of measurement, for, in the final analysis, all polarity is relative — the appearance of things depends upon where one sits.

These comments concerning polarity are quite pertinent to the successful measurement and interpretation of conditions in oscilloscope equipment for the demands of the operating sections in these instruments have caused numerous modifications in the design of power-supply circuits.

Power Transformer — Rectifier Tube Circuit Variations

One of the interesting aspects of power supplies is the variation to be found in the power transformer—rectifier tube circuit arrangements so as to satisfy not only the different voltage needs of a system, but also the different polarities of these voltages relative to ground. In order to show these most effectively, the circuits which follow will not contain the filter systems, for these have no bearing on the modifications of the balance of the circuits.

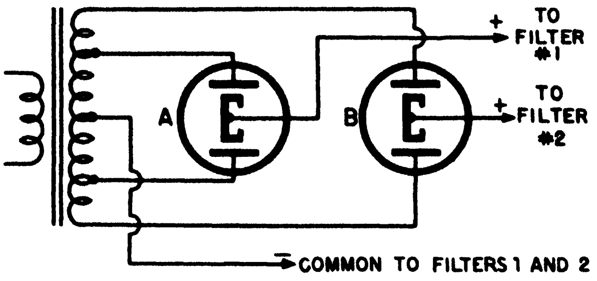


Fig. 10-133.—Single high-voltage winding tapped to serve two separate full-wave rectifying systems.

In Fig. 10-133 is shown a single winding tapped so as to serve the high-voltage requirements of two separate full-wave rectifying systems *A* and *B*, each of which demands a different a-c voltage input. When we identify this voltage as "high," it has a relative meaning; in one system, it may be high by comparison with other voltages being rectified, and in another system, these same values may be "low"voltages.

Concerning the circuits, the transformer center tap is common to both rectifying systems, and the current for the higher-voltage rectifier *B* flows through that portion of the winding which supplies the tubes receiving the lower voltage *A*, whereas the current for the low-voltage tubes flows through only that part of the winding which is common to both rectifiers. Each of the rectifier systems has its one positive output lead which connects to its own filter, whereas the negative lead from the high-voltage winding center tap is common to both filter systems. The kinds of filters which would be used with these rectifiers are matters of individual design and are not influenced by the special power-transformer design.

Another interesting arrangement is shown in Fig. 10-134. Here a single high-voltage winding serves two full-wave rectifiers; one of these *A* consists of two half-wave rectifier tubes, whereas the other *B* has both half-wave sections inside of a single envelope. The positive legs of each system are center taps of the filament windings and each connects to its individual filter system. The center tap of the high-voltage winding is the negative leg which is common to both filters.

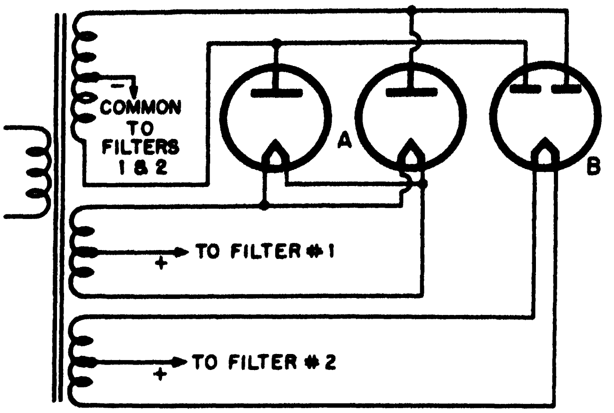


Fig. 10-134.—Two full-wave rectifiers in parallel, (A) consists of separate half-wave rectifiers, and (B) has both in the same envelope.

Another version of a dual rectification system is shown in Fig. 10-135. Here a "low" voltage rectifier of the full-wave kind delivers a voltage that is positive relative to ground, and a separate and individual "high" voltage rectifier delivers a voltage which is negative to ground. For identification purposes, the low-voltage rectifier plates are *P1-P2* and its cathode is *K*. The power transformer contains a single continuous winding which is tapped at several points. Points *a, b,* and *c* afford the voltage necessary for the low-voltage full-wave rectifier and the center tap on this winding is connected to ground. Neglecting for the moment any other possible system, what we have de-

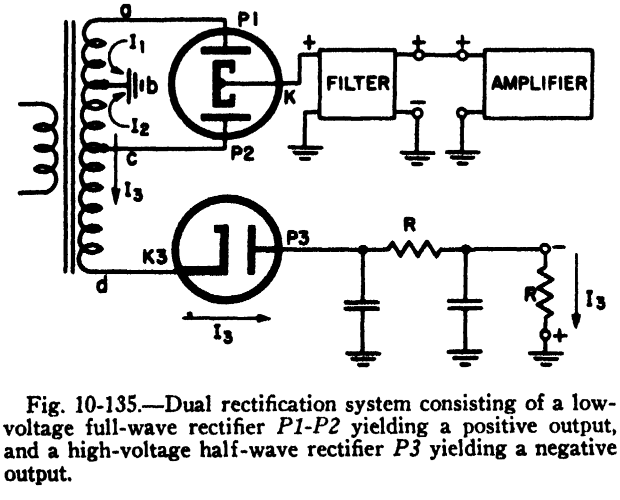


Fig. 10-135.—Dual rectification system consisting of a low-voltage full-wave rectifier *P1-P2* yielding a positive output, and a high-voltage half-wave rectifier *P3* yielding a negative output.

scribed is a conventional rectifier in which the high-voltage winding center tap is the grounded negative point. The cathode *K* common to both plates of the full-wave rectifier tube is the positive point and this feeds into a filter which we illustrate in block form. The output of the filter is shown connected to a hypothetical amplifier to which it is furnishing the plate voltage in conventional manner. There is nothing unusual about this arrangement.

Referring to the remainder of the same illustration, let us examine the extension of the previously mentioned plate winding. A voltage will exist between points *b* and *d*. It will afford a voltage which is determined by the turns ratio between this segment of the complete secondary winding and the primary winding and the primary voltage. The fact that another portion of the winding may be connected to a rectifier system is immaterial to the behavior of the second rectifier tube containing the electrodes *P3* and *K3*. Furthermore, the fact that ground point *b* on the transformer winding is negative with respect to the output of the full-wave rectifier does not necessarily mean that it must be negative for any other rectifier. Ground has a dual polarity; it may be positive as well as negative. It can accept charges and it can release charges. Its instantaneous polarity is a matter of what conditions exist in the circuit of which it is a part.

In the case shown, the secondary voltage will have a polarity which will make the ground point *b* and the point *d* alternately positive and negative. In fact, there will be correspondence between the points *c* and *d* in polarity relative to ground point *b*. When the latter is positive, points *c* and *d* are negative. When the cathode *K3* is negative relative to its plate, the tube is in a conducting state and a pulse of rectified current will pass through the half-wave rectifier, through the load resistor *R* and back to cathode *K3* via the grounded point *b*. The significant detail here is that the direction of the current through the load resistor *R* makes its upper terminal negative and its lower terminal (the one connected to ground) positive, so that we have available a negative voltage relative to ground. The half-wave rectifier is then a negative high-voltage supply. It does not interfere with the performance of the low-voltage positive polarity rectifier.

Inasmuch as every point in an a-c system may be positive at one instant and negative at the next, it is conceivable to use the same voltage point for two different kinds of power supplies by properly connecting the rectifier tubes, so that only one is operating at any one instant. Of course, both will afford output voltages. The fact that these two systems will function alternately at the respective rectifiers will not affect the output. One may be negative and the other may be positive and each will be effective by itself just as if the other one did not exist at all.

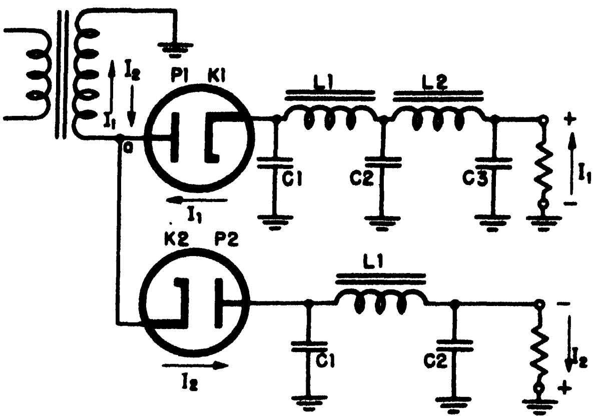


Fig. 10-136.—Dual rectification system consisting of two high-voltage half-wave rectifiers; the system *P1-K1* has a positive output and *P2-K2* a negative output.

What we have in mind is shown in Fig. 10-136. Point *a* on the high-voltage winding is common to two half-wave rectifier tubes; to the plate in one case and to the cathode in the other. The other end of the winding is grounded, which makes it common to the two rectifier outputs. One of these is a positive voltage output and the other is a negative voltage output, with respect to ground. When point *a* becomes positive relative to ground, conduction takes place between *K1-P1* and the direction of the current through the associated rectifier is shown by the arrows labeled I_1. When point *a* on the power transformer becomes negative relative to ground, conduction takes place between *K2-P2* and the current through the rectifier system is shown by the arrows labeled I_2.

An arrangement of this type could very conveniently furnish the operating potentials to the accelerating anode of a cathode-ray tube by connecting that point

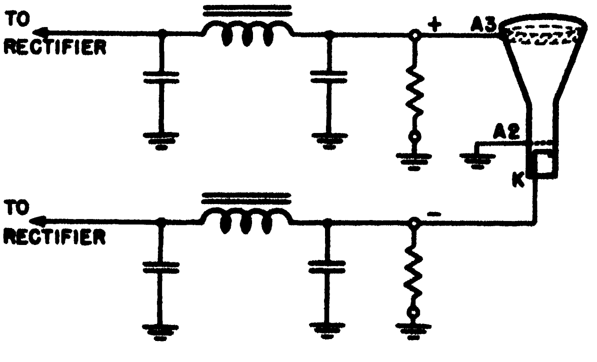


Fig. 10-137.—Application of the system of Fig. 10-136; the positive output is applied to the intensifier anode, the negative output to the cathode, and the accelerating anode is grounded.

to ground, the cathode to the negative supply, and the post-deflection intensifying anode to the positive supply, as shown in Fig. 10-137. By making the cathode negative with respect to the grounded accelerating anode, the latter is made positive relative to the cathode by an equal amount. By connecting the post-deflection anode to the positive power supply, this element is made positive relative to the grounded accelerating anode by the voltage of the positive supply, and positive relative to the cathode of the cathode-ray tube by the sum of the voltages of the two power supplies. If the negative supply is rated at −1,600 volts and the positive supply is rated at +2,000 volts, the accelerating anode is 1,600 volts positive relative to its cathode and the post-deflection intensifier ring anode is 3,600 volts positive relative to the tube cathode.

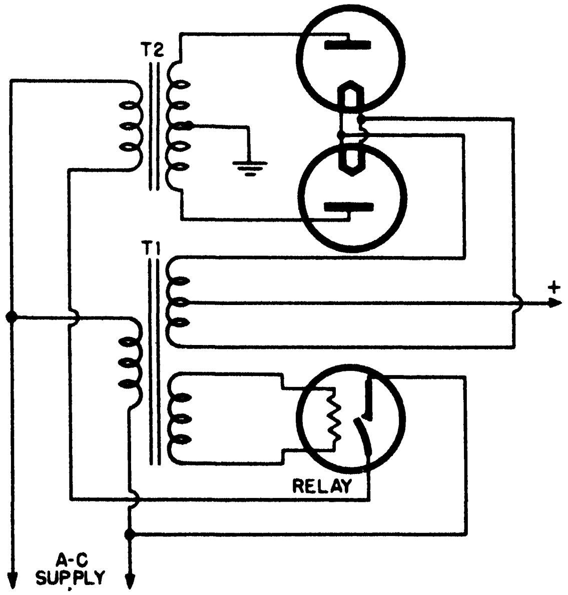


Fig. 10-138.—Illustrating the use of a time-delay relay for the delayed application of the high voltage in a mercury-vapor rectifier system.

The use of mercury vapor type rectifier tubes demands certain special transformer arrangements regardless of what other type of rectifiers may be used in the same instrument. Inasmuch as the a-c voltage to be rectified cannot be applied to the mercury vapor tubes until the heater in the device has vaporized the mercury (otherwise the tube or tubes may be damaged), it is customary to include some protective arrangement in the circuit which controls the application of the "high" a-c voltage. This may be done in a number of ways, but one commonplace system is shown in Fig. 10-138. A time-delay relay receives its heater current from the same transformer *T1* which supplies the heater current to the mercury vapor tubes and supplies heater and plate voltages to other high-vacuum rectifiers. The plate voltage for these tubes is secured from a separate transformer *T2*, the primary circuit of which contains the time-delay relay contacts as a series element.

When the primary power is applied to *T1*, only the heaters for the mercury vapor tubes are activated and the time-delay relay is heated. Depending upon the setting of the relay, the application of the plate voltage to the mercury vapor tubes is delayed by the relay until such time as the gaseous rectifiers have had a chance to become properly heated. This may involve a time lag of from 30 seconds to 1 minute. During all this time, the primary circuit of the high-voltage plate transformer *T2* is open, and it becomes closed automatically when the relay contacts are closed. Then the plate voltage is applied to this rectifier. Other rectifiers of the high-vacuum type may be part of the power-supply system but they do not require protection of this kind.

Voltage Regulators

No matter how well the filter system of a power supply may be designed, some variation in output voltage will occur as the current drain is changed. The changes in current load cannot be held constant under all conditions of equipment operation. In the final analysis, the equipment load must perform its allotted function and in so doing may undergo wide changes in current drain in accordance with the signal requirements. Yet the same equipment may demand a constant voltage. To satisfy all of these needs, recourse is had to voltage regulators as shown in Fig. 10-139. These are devices which keep the output voltage constant by acting as automatic variable voltage dropping systems. As the current drain on the supply increases or decreases, the voltage drop across the voltage regulator decreases or increases, thus keeping the power-supply output voltage at a constant level.

If the load on the rectifier is reduced so that the output voltage tends to rise, the regulator element in series with the load automatically increases its resistance and thus increases the voltage drop across it, thus keeping the output voltage constant. If the load on the power

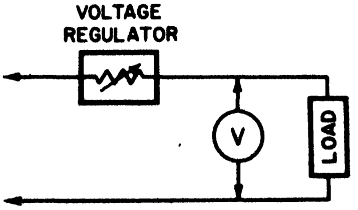


Fig. 10-139.—Block diagram of an elementary voltage regulator system.

supply is increased, tending to reduce its output voltage, the regulator element would automatically decrease its resistance, thus raising the output voltage to its previous level, and keeping it constant.

Naturally the regulator must be designed for operation over a specific range and the vacuum tubes used for this purpose bear definite ratings of current and voltage which must be adhered to, if regulation is to be attained. Sometimes instead of a simple tube which can accomplish this action, a complicated amplifier system is used.

One of the most frequently used voltage regulators is the "glow" tube, a gaseous discharge tube across whose terminals the voltage drop will remain constant over a wide range of currents through the tube. This action stems from the latitude in the degree of ionization of the gas within the tube. If the current through the tube is high, the ionization is great, whereas if the current through the tube is small, the degree of ionization of the gas is small, while a fixed voltage drop is maintained across the tube. Understandably, the tube bears a rating of minimum current in order that it be effective. If this cannot be attained, then the tube will not function because the gas will not be ionized. Whatever this minimum requirement may be is naturally taken care of in the design of the power supply. At the same time, each tube also bears a rating of maximum current which may be passed through the device without damaging it. These data are stated in the type numbers assigned to such units.

Examples of these tubes are the VR-75, the VR-105, and the VR-150. The first two letters indicate that it is a voltage regulator. The numbers 75, 105, and 150 state the constant voltage which appears across the terminals of the tube for rated current drains. One important point which must be borne in mind is that the voltage reference dictates that at least that much voltage must be available across the terminals of the tube in order that the tube "fire" or the gas be ionized. For example, a simple application is shown in Fig. 10-140A. If this tube is rated at 75 volts, it means at *at least* 75 volts must be available across points *a-b* to start the action. If this condition is satisfied, then the voltage output will be kept at 75 volts for a wide variety of current drains. However, if for some reason the voltage across *a-b* is allowed to fall too far below the regulator tube rating, the tube will cease firing and the regulating action will cease.

An example of the location of such a glow discharge voltage regulator in a power-supply system is shown in Fig. 10-140B. Load *A* is the main load on the power supply and does not receive a regulated voltage even though the voltage regulator is active in the system. Load *B*, on the other hand, receives a regulated voltage because it is the voltage across the regulator tube. It is conceivable that the output voltage from the supply as a whole is within the control of a single regulator tube, in which case, the circuit would be as in *A* rather than as in *B*.

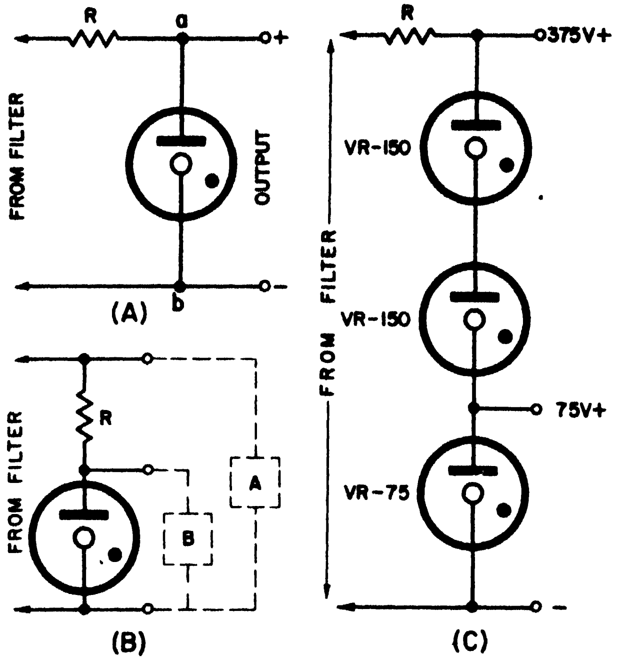


Fig. 10-140.—Various arrangements of glow discharge voltage regulator tubes; (A) is a simple 75-volt regulated output, (B) has a regulated output and an unregulated output at a higher voltage, and (C) is a regulated supply having 75- and 375-volt outputs.

Although the voltage ratings of these tubes seem limited, their performance is such that they may be used in various series combinations so as to accommodate a variety of voltage requirements. Thus three VR-75 tubes may be used in series so as to regulate 225 volts, or the same regulation can be secured by using a VR-75 in series with a VR-150, or three VR-150 tubes may be connected in series to regulate 450 volts, etc. Such an arrangement is shown in Fig. 10-140C. This system affords a regulated 375 volts and a regulated 75 volts.

Another aspect of these devices is their behavior relative to the removal of ripple. A regulator tube tends to keep the voltage constant and a ripple is tantamount to a variation in voltage. In effect, the use of such voltage regulators is the equivalent of having added capacitance to the filter section, in that way reducing the "hum" in the output and obtaining a more constant output voltage. It also is beneficial as a means of attaining a low impedance at the output of the power

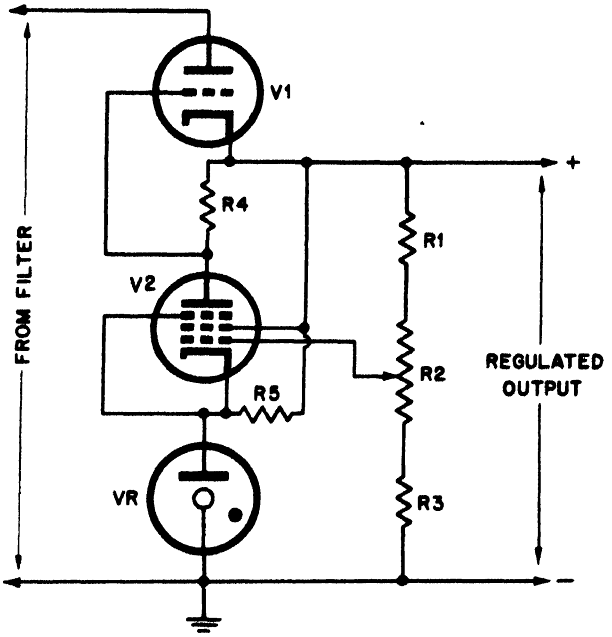


Fig. 10-141.—Schematic of a basic electronic voltage regulator system.

supply so that the possibility of coupling between different sections of the oscilloscope via the common impedance of the power-supply output circuit is minimized.

Voltage regulators of more complex nature also are used. These are used when the current requirements exceed the capabilities of the glow discharge tubes. A fundamental system of this kind is shown in Fig. 10-141 and the specific arrangements used in some oscilloscopes are elaborations of this basic circuit. Since the discussion that accompanies the schematic wiring diagrams of the commercial devices explains the circuits used in the instruments, this explanation of a basic system might serve as a foundation.

The regulated output voltage appears across the series combination of *R1*, *R2*, and *R3*, although these three resistors do not comprise the complete network. A part of it is tube *V1*, which is operated as an automatically varying resistance which, in effect, is in series with the three resistors previously mentioned. Thus, by varying one of the elements automatically, the voltage across the other three is maintained constant over a wide range of currents. These flow through *V1* and are controlled by the pentode tube *V2*.

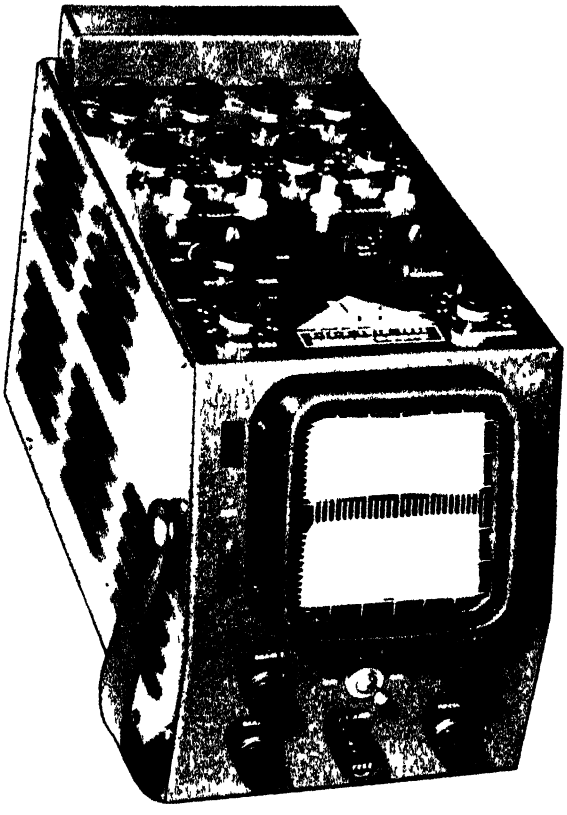
In order to follow the action of this circuit, let us set up a number of conditions. The cathode of *V1* has a bias resistor *R4*, which is common to the plate circuit of *V2*, and which we can call the load resistor for that plate circuit. The plate of *V1* derives its plate voltage from the unregulated, but filtered, supply. The grid of that tube receives a bias from the bias resistor *R4*, and, as can be seen, the plate-cathode resistance of tube *V1* is the series element between the unregulated and the regulated positive leads of the system. By variation of the grid bias of *V1*, control of the plate-cathode resistance of that circuit is attained. This bias is controlled in turn by the bias conditions in *V2*, since the plate current of that tube flows through the cathode resistor of *V1*. The bias conditions in *V2*, on the other hand, are controlled by the setting of the potentiometer *R2* so that the potential at the grid of *V2* is less positive than its cathode by whatever amount is necessary to pass that amount of plate current which will result in the required voltage across the positive-negative output terminals. This is the voltage the system is supposed to maintain.

Let us assume that the load voltage tends to rise because of reduced current drain, or due to an increase in voltage from the filter system. For the moment, until regulation sets in, the voltage across the *R1-R2-R3* network also tends to rise so that the voltage applied to the control grid of *V2* tends to make that grid less negative. The cathode voltage being controlled by the glow tube tends to remain constant. Since the control grid has become less negative, the plate current in *V2* tends to increase. This, in turn, is the equivalent of an increase in current through *R4*. Therefore, the control grid of *V1* becomes more negative and increases the plate-cathode resistance of *V1*, causing an increased voltage drop across that resistance and tending to reduce the voltage across *R1-R2-R3* to its original value.

If the change in voltage were in the direction opposite to that mentioned, the control grid of *V2* would become more negative, thus reducing the bias voltage on the control grid of *V1*, since the current through the cathode resistor *R4* would be decreased. This would lower the plate resistance of *V1* and thus reduce the voltage drop across that tube, tending to maintain the voltage across the output.

Multibeam Oscilloscopes

The multibeam oscilloscope is a variation as far as basic oscilloscopes are concerned in that it may employ a cathode-ray tube which contains two or more guns or it may consist of two individual single-gun cathode-ray tubes. This is an innovation in multiple trace oscilloscopes, both as to the use of two individual tubes and also as to the rectangular shape of the screens. An example of such a device is shown in Fig. 10-142. (The multibeam oscilloscopes are shown in Fig. 1-14 in Chapter 1.) In Fig. 10-142, the two screens are located parallel to each other with a slight separation between



Courtesy Waterman Co.

Fig. 10-142.—Multibeam oscilloscope containing two individual single-gun cathode-ray tubes.

them as indicated by limits of the wide areas on the front panel of the instrument.

Regardless of whether the device consists of a multigun cathode-ray tube — which, as we have said earlier in this book, may contain as many as ten guns, thus affording ten individual beams — or it consists of two individual tubes, each with its gun assembly and its own deflection system, each tube or gun has its own individual sets of beam controls, amplifiers, and the like.

The manipulation of these controls is in accordance with what has already been said for the basic oscilloscope, and requires no elaboration. Relative to the individual traces which may be developed, all of the facts given in this chapter apply to the development of the trace, the amplitude control, the relationship between the frequency of the signal being observed and the frequency of the sweep, etc. If anything is singular, it is found in some devices wherein a single sweep may be used for the production of the horizontal dimension for all of the horizontal-deflection systems. This is a facility which is furnished, even though identical time bases are provided for each beam so that each gun and deflection system may be utilized independently of the other for whatever display is desired. Of course, in the case of the multigun cathode-ray tube, only a single screen is available, inasmuch as all of the guns are contained within a single envelope. In the case of the dual tube arrangement, individual screens are provided for the individual displays.

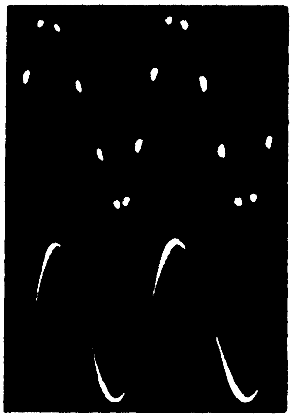


Fig. 10-143. — Oscillogram showing intensity modulation by signals of different frequencies. At the top, a 60-cps signal is intensity-modulated at 480 cps. At the bottom, the same 60-cps signal is modulated at 120 cps.

Relative to the manipulation of the controls so as to position the beams or the traces, those devices which contain the multigun single-screen type of tube afford the means whereby the traces controlled by the different amplifiers may be made to overlap so as to enable comparisons of various kinds. Examples of these are shown in Figs. 10-143 through 10-146. In Fig. 10-143 are shown two waves which are slightly displaced horizontally relative to each other. One of these is intensity-modulated at eight times the signal frequency and the lower one is intensity-modulated at twice the signal frequency. In Fig. 10-144 are shown two waves which have been superimposed along the horizontal axis so as to illustrate the phase relationships between them. In Fig. 10-145 is shown an extremely interesting application of a dual-beam oscilloscope. One channel develops the complete pattern of a phenomenon as shown in the upper trace. The lower trace is for the same phenomenon on the second channel except that this channel, utilizing a much faster sweep, spreads a portion of the upper trace, thus enabling easier analysis. This is an illustration of the advantage of having individual sweeps for each of the guns in the tube. In

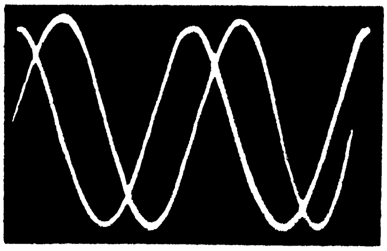


Fig. 10-144.—Oscillogram showing phase comparison between two sine waves. They are plotted on a common sweep and superimposed to facilitate measurement.

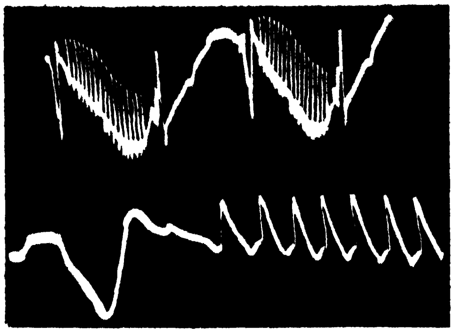


Fig. 10-145.—A complex waveform taken from an ordinary fluorescent lamp (top) is compared with a greatly expanded portion of the same waveform (bottom). Expansion is accomplished by plotting the lower trace on a much faster sweep.

Fig. 10-146 is shown the action of a differentiating circuit acting on a square wave. One beam and its amplifiers act on the square-wave signal at the input of the differentiating circuit. The action is displayed by the upper trace. The lower trace shows the same signal after differentiation. Both signals are plotted on a common sweep, which means that the same sweep was used to drive the two beams in the horizontal direction, whereas individual vertical amplifiers were used to feed the two signals to the respective deflection systems.

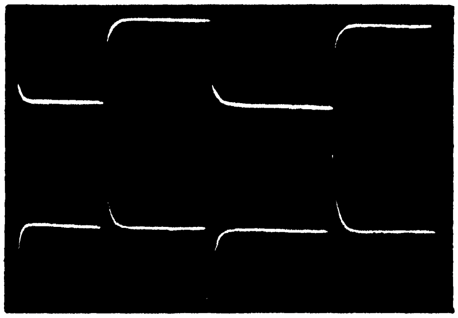


Fig. 10-146.—Oscillogram showing a square-wave signal (top) and the resultant waveform after differentiation by an *R-C* network (bottom). Both signals are plotted on a common sweep.

All of the traces shown in Figs. 10-143, 10-145, and 10-146, which are displaced horizontally, may be positioned so that they are superimposed to facilitate studying by the proper manipulation of the positioning controls associated with each of the channels. Such superpositioning is not possible in the device which employs two individual tubes. Comparison is nevertheless possible since the two tube screens are parallel and fairly close together. Thus, the traces shown in Figs. 10-143, 10-145, and 10-146 could well appear on the two-tube unit with the upper trace on one tube screen and the lower trace on the other tube screen. In all of these devices, each system or each channel is in reality an independent oscilloscope and the handling of the controls may be carried out on that basis. The ability to use the same sweep for both tubes or for the two sets of deflection systems is an added feature for purposes of flexibility.

Relative to the nomenclature assigned to controls on these devices, they are labeled in the same manner as the single-gun or single-tube instruments, and it is customary to group the controls for each channel separately. In this connection, the various controls which function on the beam may bear the label BEAM A, BEAM B, or BEAM 1 and BEAM 2. Relative to the amplifier, sweep, sync, and similar controls, they may be identified as CHANNEL A and CHANNEL B, or as CHANNEL 1 and CHANNEL 2, etc.

Concerning the performance characteristics of the section which comprise the different channels, invariably they are identical both as to frequency response and as to sensitivity. However, this deserves some qualification, for, as we described in Chapter 8, a certain tolerance is employed in the manufacture of the guns and the deflection systems so that perfect coincidence of two traces may not always be possible. However, the deviation in the case of the multigun tubes is never so great as to impair the utility. As a rule, manufacturers try to compensate in their amplifier systems for such tube tolerances. Of course, this is no problem in the multitube oscilloscope, since the two traces cannot be superimposed.

FREQUENCY COMPENSATION

Although this book is not intended as a text on the design of oscilloscope amplifiers, it seems fitting to deal with the matter of low- and high-frequency compensation in more than casual terms. We have placed so much emphasis on the matter of frequency response in this chapter that a short discussion on the means of attaining the necessary response will not be out of place.

The mathematical analysis of frequency compensation may be neglected by the individual who has no

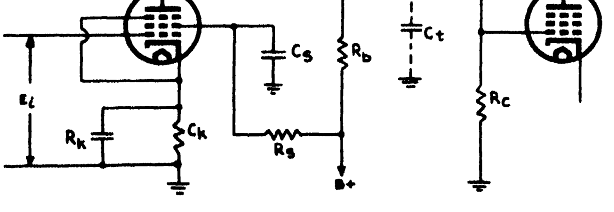


Fig. 10-147.—Typical *R-C* coupled amplifier.

interest in such details without destroying the intent of the discussion.

The basic circuit for a vertical or a horizontal amplifier is an *R-C* coupled amplifier as shown in Fig. 10-147. This circuit has no compensation for either the high- or low-frequency response. The gain will fall off at both the high- and low-frequency ends. Let us briefly discuss the reasons for the response dropping off.

Considering the low frequencies, we can see that the coupling capacitor C_b is in series with the grid resistor R_c of the following tube. The output signal from the first tube is applied across this series combination. As the frequency becomes lower, the reactance of the capacitor C_b becomes higher, and less of the output signal appears across resistor R_c to be applied to the next tube. There is also a phase shift due to this resistor-capacitor combination.

At the high frequencies, the effect of C_t becomes appreciable. C_t represents the total shunt capacitance due to the output capacitance of the first tube, the input capacitance of the second tube, and all the stray capacitance due to the wiring. At the high frequencies, the reactance of this capacitance becomes small enough to lower the impedance of the plate load of *V1*. The plate load, instead of being a pure resistance, will then be a resistance in parallel with a capacitive reactance. This smaller load impedance will cause the output voltage to drop and the capacitive component of the load will cause a phase shift in the output signal.

Because of the deficiencies mentioned above, it is evident that the ordinary *R-C* coupled amplifier is not good enough for use as a vertical amplifier without some form of compensation to offset the falling off in response at the low and high frequencies and to give a linear phase shift to prevent phase distortion. We will now consider the different methods used for low- and high-frequency compensation.

Low-Frequency Compensation

We have seen how the combination of coupling capacitor C_b and grid-leak resistor R_o reduced the low-frequency response of the vertical amplifier. These effects of attenuating the low frequencies must be compensated for in wide-band amplifiers. These amplifiers employ pentode tubes of such design that they have relatively high values of mutual conductance g_m and small values of interelectrode capacitances. The reasons for these design features will soon be evident.

When an *R-C* coupled amplifier is compensated at the low-frequency end, another resistance R'_b is placed in series with the plate-load resistor R_b and a capacitor is placed across the added resistor. Fig. 10-148 shows a typical low-frequency compensated amplifier, in which $R'_bC'_b$ is added to better the low-frequency response. This addition of $R'_bC'_b$ appears to be nothing more than a decoupling network. In reality it is a decoupling network with the screen *B*+ lead connected directly to *B*+ instead of between R_b and R'_b. Besides its immediate use of raising the low-frequency response, it decouples any undesired feedback voltage due to the common impedance of the plate supply.

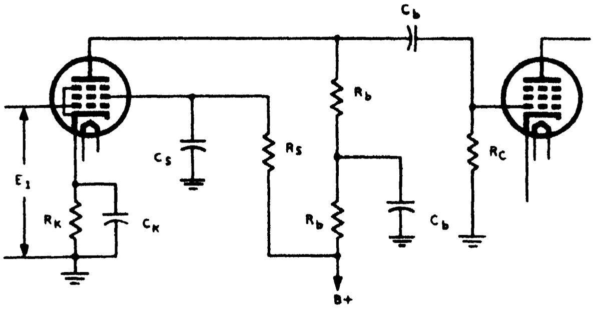


Fig. 10-148—Typical low-frequency compensated *R-C* coupled amplifier. The decoupling network $R'_bC'_b$ has been added to the circuit of Fig. 10-147 to better the low-frequency response.

Since the tube used is a pentode, we can make some approximations for simplifying the analysis of this circuit. The grid-leak resistor R_c is much larger than the plate-load resistor R_b and the plate resistance of the tube r_p is much greater than R_b; therefore, all that is necessary is to take into consideration the essential effect of the combination R_cC_b. At low frequencies, R_cC_b causes a drop in signal potential from the plate of tube *V1* to the grid of tube *V2* and also causes phase shift between the two tubes. The coupling capacitor C_b causes the shift. Why, then, is the value of C_b not made large enough so that it will offer negligible reactance at low frequencies? The value of C_b could be increased and this would improve the low-frequency response, but since we are dealing with frequencies of several hundred kilocycles and higher, the larger the value of C_b, the greater will be the shunt capacitance to ground that it introduces. These shunt capacitances, although extremely small, are very detrimental at the high frequencies. Low-frequency compensation is used to avoid the introduction of these shunt capacitances. The insertion of $R'_bC'_b$ compensates for the loss in gain at the low frequencies and also introduces a phase shift to compensate for the undesired phase shift introduced by R_cC_b.

Since the combination of R_cC_b at low frequencies reduces the gain, the combination of $R'_bC'_b$ at these same frequencies is inserted to offer a higher load impedance to tube *V1*, and increase the gain. Increasing the load impedance will increase the effective voltage across the load and bring the gain back to normal. Again referring to Fig. 10-148, the operation is as follows. The voltage across the grid-leak resistor decreases (thereby decreasing the input to the grid of tube *V2*) because the reactance of C_b increases as the frequencies are decreased. The reactance of capacitor C'_b increases at the same time. As a result, the parallel combination of $R'_bC'_b$ is seen to be an impedance that is added to the plate-load resistor R_b and effectively increases the total load impedance on the tube *V1* as the frequencies are decreased.

Since this amplifier stage uses a pentode and $r_p > R_b$, the gain at low frequencies in the case of this compensated stage is expressed as

$$A_{\text{low}} = A_{\text{int}} \frac{1}{\sqrt{1+\left(\dfrac{X_{Cb}}{Z_{\text{load}}}\right)^2}} \qquad (10\text{-}13)$$

where

A_{low} = the gain at low frequencies
A_{int} = the gain at intermediate frequencies
X_{Cb} = the reactance of the capacitor C_b
Z_{load} = the total effective load impedance, including R_b, R'_b, and C'_b.

From this equation it may be seen that the effect of the original plate-load resistor R_b has been changed to Z_{load}. Since the plate-load impedance has been increased by the $R'_bC'_b$ combination, the gain at low frequencies is also increased. As the frequency decreases, the ratio of X_{Cb}/Z_{load} also decreases, making equation (10-13) approach the gain at intermediate frequencies, A_{int}. It would be ideal if resistor R'_b could be increased to an infinite amount but the need for an appreciable amount of plate voltage on the tube makes this impossible. If this could be achieved, the gain would be equal to that of the tube itself and no loss would be encountered; the gain would be maximum and flat throughout the response. From *experimentation* it has been found that, for the proper response, the time constants of $R'_bC'_b$ and R_cC_b should be equal to each other. The voltage loss due to C_b will then be compensated (in direct proportion) by the voltage gain due to the total load impedance, in such a way that the voltage across the grid-leak resistor R_c will remain constant. In many instances, the combination of $R_bC'_b$ is chosen first and the coupling capacitor C_b is made as large as is desired, then the grid-leak resistor R_c is determined. For instance, assuming that C'_b is chosen to be 40 μf, C_b chosen as 0.25 μf, and the time constant of R_cC_b to be 0.1 second (100,000 microseconds) it is desired to know what the values of R_b, R'_b, and R_c are at the lowest frequency of operation, which in this case is assumed to be 5 cps. Therefore

$$R_cC_b = R_b\,C'_b = 100{,}000\ \mu\text{sec}$$

and since $C'_b = 40\ \mu f$,

$$R_b = \frac{100{,}000}{C'_b} = \frac{100{,}000}{40} = 2{,}500 \text{ ohms.}$$

Since $C_b = 0.25\ \mu f$

$$R_c = \frac{100{,}000}{0.25} = 400{,}000 \text{ ohms.}$$

R'_b should be chosen to be about 20 times greater than the reactance of C'_b at the lowest frequency to be passed (5 cps in this case) and yet not drop too much of the supply voltage. Therefore

$$R'_b = 20X_{C'b} = \frac{20}{2\pi fC'_b}$$
$$= \frac{20}{(2\pi)(5)(40\times10^{-6})} = 15{,}900 \text{ ohms.}$$

We, therefore, find that

R_b —— should be about 2,500 ohms
R_c —— should be about 400,000 ohms
R'_b —— should be about 16,000 ohms.

The equation for the gain at low frequencies of an uncompensated *R-C* coupled amplifier is

$$A_{\text{low}} = A_{\text{int}} \frac{1}{\sqrt{1+\left(\dfrac{X_{Cb}}{R_c}\right)^2}} \qquad (10\text{-}14)$$

If this equation is studied, it will be found that the phase shift, due to the R_cC_b combination, at the low frequencies is

$$\theta = \tan^{-1}\frac{X_{Cb}}{R_c} \qquad (10\text{-}15)$$

where θ is in degrees. It can be seen that the coupling capacitor C_b causes a lead in phase. Inserting the $C'_bR'_b$ network into the load tends to cancel this phase lead by introducing a phase shift in a lagging direction.

Besides the grid-leak resistor and coupling capacitor combination, there are two other resistance-capacitance combinations that cause a falling off in gain and a leading phase shift at the low frequencies. First, there is the effect of the R_kC_k bias impedance. The best way to eliminate this effect would be to eliminate R_kC_k and apply bias another way. This is usually not desired, as cathode bias is preferable in many cases, so it has been determined experimentally that this undesired effect can be compensated for by making the time constant of R_kC_k equal to the time constant of $R'_bC'_b$. Mathematically we have

$$R_kC_k = R'_bC'_b \qquad (10\text{-}16)$$

Secondly, there is the effect of the screen-grid circuit. A pentode tube possesses a dynamic screen-grid resistance as well as a dynamic plate resistance. This screen-grid r_{SG} in combination with the bypass capacitor C_S also causes a falling off in response and a phase shift (leading) at the low frequencies. Since it is difficult to vary such an element as r_{SG}, and because C_S is the only variable parameter, little compensation can be accomplished by a single stage; better compensation is obtained by the use of several stages of amplification.

A very important thing to note is that in all these compensating problems $R'_bC'_b$ plays the important role. Although $R'_bC'_b$ can be used to compensate for any one of the above cases, only one case can be compensated for in a stage such as shown in Fig. 10-148. A usual practice in cathode-biased amplifiers is to make R_c very large (this is usually possible) and, therefore, the time constant of R_cC_b will be large and the needed compensation is shifted to the R_kC_k combination (instead of the $R_cC_b = R_bC'_b$ combination).

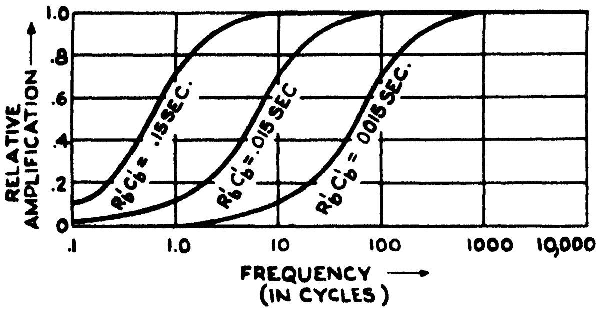


Fig. 10-149.—The effect of inserting different values of the $R'_bC'_b$ network is shown. As the time constant of the decoupling network is increased, the low-frequency response increases.

In Fig. 10-149 are several curves showing the effect of the insertion of the $R'_bC'_b$ network on the low-frequency response for different values of $R'_bC'_b$. From these curves it is seen that as the time constant of the $R'_bC'_b$ combination is increased, the low-frequency response increases. However, there are practical limitations to the values of R'_b and C'_b that can be chosen.

High-Frequency Compensation

High-frequency compensation is even more important than the low-frequency compensation because of the higher band of frequencies to be passed. In oscilloscope amplifiers, frequencies as high as several hundred kilocycles and even several megacycles have to be passed with a fairly flat response from the minimum frequencies on upward. Therefore, this type of compensation is considered as the more important one.

There is really only one part of the R-C coupling circuit that reduces the response at high frequencies, as compared with the three different networks affecting the low frequencies. Referring to Fig. 10-147, the total shunt capacitances, represented by the capacitance C_t, existing in the R-C circuit are the chief reasons for the high-frequency response falling off.

Since these shunting capacitances are difficult to reduce due to their inherent qualities, compensation must be made to offset their effects. Since we are dealing with high frequencies, the reactance of these shunting capacitances is very low and offers a low-impedance path to these frequencies. Considering a single coupling stage, these capacitances include the output and input capacitances of the first and second tubes, respectively, the interelectrode capacitances of the tubes themselves, and any stray wiring capacitances, such as those between the coupling capacitor C_b and ground. The equation for the gain at high frequencies, assuming $r_p > R_b$ and $R_c > R_b$, becomes

$$A_{\text{high}} = \frac{g_mR_b}{\sqrt{1 + \left(\dfrac{R_b}{X_{Ct}}\right)^2}} = \frac{g_mR_b}{\sqrt{1 + (2\pi fC_tR_b)^2}}. \qquad (10\text{-}17)$$

This relation states that the gain of an R-C coupled amplifier stage at high frequencies A_{high} is dependent upon four distinct factors: the transconductance or g_m of the tube used; the plate load, or R_b, on the same tube; the frequency of operation f; and the total shunting capacitances C_t. Consequently, from the above expression, it is found that if the g_m of the tube in question is high, so will the gain A_{high} be high; but if the shunt capacitances, C_t, of the tube and the frequency of operation f are high, the gain will decrease.

Assuming R_b equals 50,000 ohms, g_m equals 1,500 micromhos, f equals 1 Mc, and C_t equals 30 $\mu\mu f$, then

$$A_{\text{high}} = \frac{(1500 \times 10^{-6})(50{,}000)}{\sqrt{1 + [(2\pi)(1 \times 10^6)(30 \times 10^{-12})(50{,}000)]^2}}$$

$$= \frac{75}{9.46} = 7.9$$

From the above mathematical computation, the gain at the middle frequencies, which is equal to g_mR_b is 75, as compared to the gain at the high-frequency end of the band, which is 7.9. The drop in gain, as mentioned before, is due to C_t.

Shunt Peaking

It is seen that with a tube of good mutual conductance and a circuit with the total shunt capacitances small, the gain at high frequencies will be high. The g_m of the tube is controlled by using special high-g_m pentodes designed especially for wide-band amplifiers. The undesired effect of C_t is reduced by inserting inductances in the circuit at special places. These inductances effectively increase the load impedance on the amplifier, thereby increasing the gain lost by the shunting capacitances. This is analogous to low-frequency compensation where the effective load impedance is also increased. The simplest type of compensation at the high frequencies is to insert an inductance L_p in series with the plate-load resistor R_b. L_p is known as a *peaking coil* because it effectively "peaks" the high-frequency end of the response curve. This type of stage is sometimes referred to as a "shunt-peaked" stage because the coil is placed in shunt with the tube (effectively from plate to ground).

Fig. 10-150A shows a typical wide-band oscilloscope amplifier compensated at the high-frequency end only and its equivalent circuit at these frequencies. In Fig. 10-150B the coupling capacitor C_b is omitted because at these high frequencies its reactance is so low that it becomes effectively a short circuit. The total load impedance on tube $V1$ is the combination of R_b in series with L_p, both in parallel with the shunting capacitances C_t as shown in the dotted portion of Fig. 10-150B. An analysis of this circuit is necessary in order to show how the effective load impedance is increased in order to increase the gain. (The grid-leak resistor R_c is neglected as part of the load impedance because it is usually very high and it has little effect on the parallel branch.)

The total load impedance represented by Z_L is given as follows

$$Z_L = \frac{X_{ct}\sqrt{R_b^2 + X_{Lp}^2}}{\sqrt{R_b^2 + (X_{Lp} - X_{ct})^2}}. \qquad (10\text{-}18)$$

It is known that a parallel resonant circuit has a very high impedance at resonance and, therefore, a large voltage will be developed across it. Consequently, since the effective load impedance has to be increased to off-

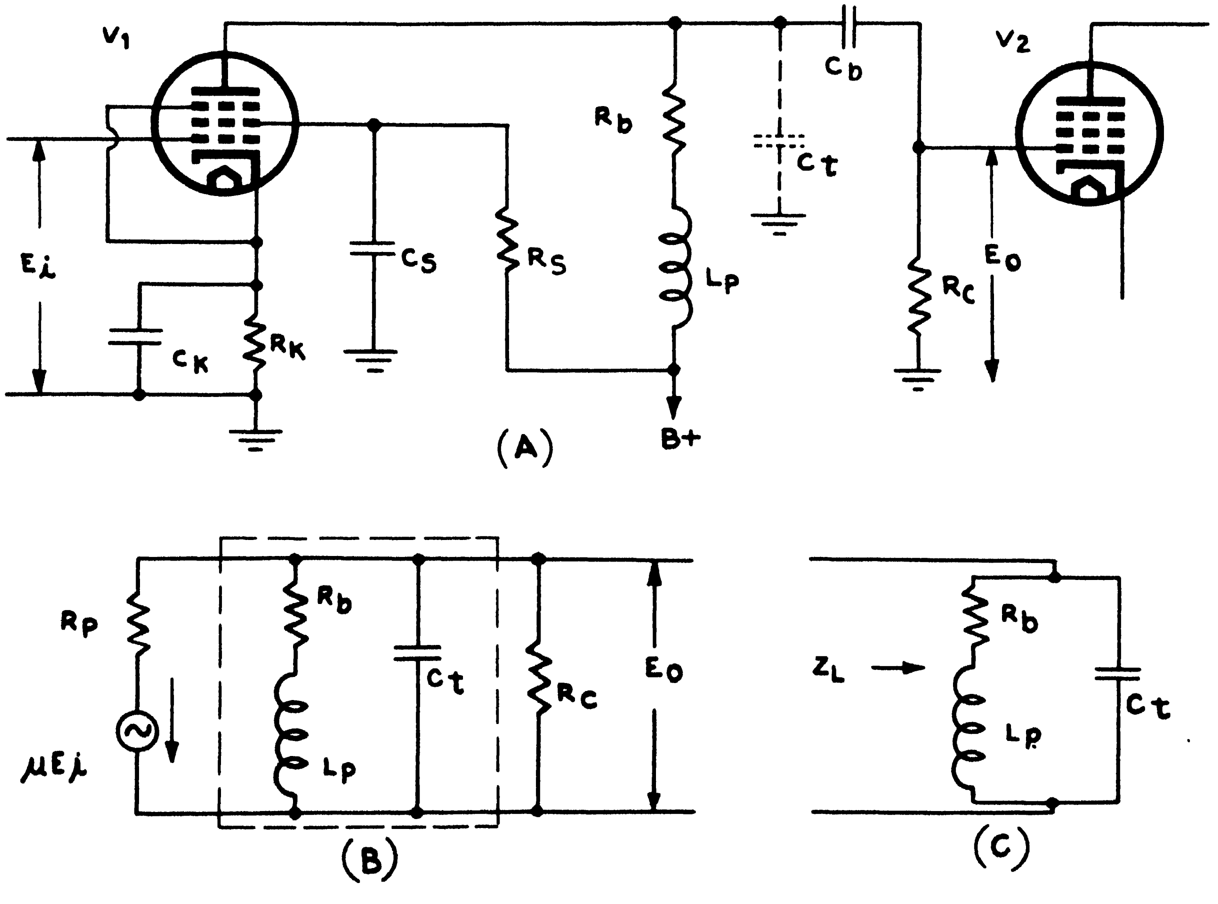


Fig. 10-150.—A typical video amplifier compensated only at the high-frequency end is shown in (A) and its equivalent circuits in (B) and (C).

set the drop in response, the parallel circuit of R_b, L_p, and C_t is made to resonate at the highest frequency that we desire to amplify. The proper values needed for this resonance are chosen from R_b and L_p, because C_t is fixed. Assuming that the resonant frequency has already been established in the circuit, the capactive reactance is equal to the inductive reactance ($X_{Lp} = X_{Ct}$), provided that the resistance R_b has a minor effect. The load impedance becomes

$$Z_L = X_{Ct}\sqrt{1 + (X_{Lp}/R_b)^2} \qquad (10\text{-}19A)$$

$$\text{or} \quad Z_L = X_{Ct}\sqrt{1 + (X_{Ct}/R_b)^2} \qquad (10\text{-}19B)$$

$$\text{or} \quad Z_L = X_{Lp}\sqrt{1 + (X_{Ct}/R_b)^2} \qquad (10\text{-}19C)$$

$$\text{or} \quad Z_L = X_{Lp}\sqrt{1 + (X_{Lp}/R_b)^2}. \qquad (10\text{-}19D)$$

Equations (10-19A) through (10-19D) are all equal because they are evaluated at the resonant frequency where $X_{Lp} = X_{Ct}$ under the above assumption.

From our previous knowledge of resonance curves, it is known that the critical frequency or the minimum frequency (for either the high or low end) acceptable for design purposes is that point on the curve where there is a 3 db drop in gain. At the high end, it is this frequency that we start with in correcting the gain because all the frequencies between the intermediate ones and this high one fall into the acceptable area. For good response in *R-C* coupled amplifiers, the load resistance is made equal to the reactance of its shunting capacitances which is what we have at the 3-db point. Therefore, we get

$$R_b = X_{Ct} = \frac{1}{2\pi f_c C_t} \qquad (10\text{-}20)$$

where f_c = the frequency where correction starts.

Under this criterion the radical of equation (10-19) becomes equal to the square root of two and the minimum load impedance that will be presented to the amplifier at the resonant frequency of L_p and C_t is

$$Z_L = X_{Ct}\sqrt{2} = 1.41\ X_{Ct} \qquad (10\text{-}21A)$$

$$\text{or} \quad Z_L = X_{Lp}\sqrt{2} = 1.41\ X_{Lp}. \qquad (10\text{-}21B)$$

The maximum frequency that has to be compensated for must be found. Assuming this frequency to be the 1 Mc mentioned, then it is known as the correction frequency for which the wide-band amplifier must be designed. When the parallel impedance becomes effectively a pure resistive load to the tube (at resonance), then the total reactance of the circuit is zero and the reactance of one branch is equal to that of the other. At this point, the inductance L_p becomes

$$L_p = C_t(R_b)^2. \qquad (10\text{-}22)$$

Thus, if we know the load resistance R_b used and have figured the shunting capacitances, we can calculate the value of shunt-peaking inductance L_p needed. It has been found from practice that when a response curve is plotted for all values of frequency versus gain, using the compensated circuit of Fig. 10-150A and equation (10-22), the high-frequency end of the curve is improved considerably over that of an ordinary *R-C* coupled circuit. The amplitude or the gain of the response curve receives the improvement, but the over-all curve is not as flat as we would desire because the value of inductance used seems to be too large and has caused too high a rise in gain at the critical frequency. The rise or peak of the curve due to coil L_p is the reason L_p is termed a "peaking" coil. From experimentation, it has been found that for over-all flat response, down to the frequency where the load resistance R_b is equal to the capacitive reactance of C_t, the peaking coil should have a value equal to *half* of that represented by equation (10-22), or

$$L_p = \tfrac{1}{2}C_t(R_b)^2. \qquad (10\text{-}23)$$

From equation (10-20), $C_t = 1/2\pi f_c R_b$; when substituted in equation (10-23), we get

$$L_p = \frac{0.5R_b}{2\pi f_c}. \qquad (10\text{-}24)$$

Equations (10-23) and (10-24) are design formulas that have actually been derived from practice. The response curve will be flat down to the frequency of correction if the above formulas are used. The actual resonant frequency of the parallel circuit is approximately $\sqrt{2}$, or 1.41 times the frequency of correction. In summarizing the above equations, we can say that the shunt-peaking inductance L_p can be determined in one of two ways:

First, all that would have to be known to find L_p would be the plate-load resistance R_b and the frequency of correction and not the shunting capacitances. This is evident from equation (10-24).

Secondly, we do not have to know the frequency of correction in order to find the shunt-peaking inductance L_p, but we would have to know the load resistance R_b and the shunting capacitances C_t. This is evident from equation (10-23).

As an example, suppose that the shunting capacitances of a stage such as that shown in Fig. 10-150A are found to be equal to 30 $\mu\mu f$ measured at a frequency of 1 Mc. We must now determine the value of the shunt-peaking coil L_p and the value of the plate-load resistor R_b needed. We know that for flat response at the high frequencies, the plate-load resistor R_b must equal the reactance of the shunting capacitances at the maximum frequency to be amplified (which then becomes the correction frequency). Therefore, with $f_c =$ 1 Mc

$$R_b = \frac{1}{2\pi f_c C_t} = \frac{1}{2\pi(1 \times 10^6)(30 \times 10^{-12})}$$
$$= 5{,}310 \text{ ohms},$$

and from equation (10-23)

$$L_p = \tfrac{1}{2}C_t(R_b)^2$$
$$= \tfrac{1}{2}(30 \times 10^{-12})(5310)^2$$
$$= 423\ \mu\text{h}$$

or from equation (10-24)

$$L_p = \frac{0.5R_b}{2\pi f_c} = \frac{0.5 \times 5310}{2\pi(1 \times 10^{-6})} = 423\ \mu\text{h}.$$

As an illustration of how the peaking inductance L_p affects the high frequencies, let us examine some typical response curves. Fig. 10-151 shows a number of different curves, where each one utilizes a different value of L_p. Note that the curve where the response is the flattest is when $L_p = \tfrac{1}{2}C_t(R_b)^2$.

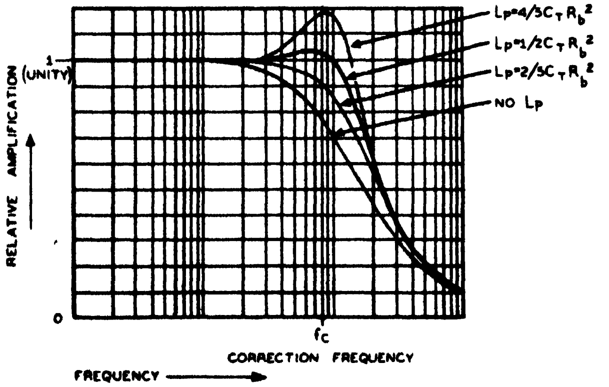


Fig. 10-151.—The manner in which the peaking inductance affects the high frequencies is shown in these typical response curves, each utilizing a different value of L_p.

Series Peaking

Another type of high-frequency compensation is known as series peaking. This type of circuit is shown in Fig. 10-152B. Here the peaking coil is placed in series with the output of the amplifier, instead of in parallel with it. This type of compensation is used when higher gain and more linear phase shift are desired than a shunt-compensated circuit provides. (In the previous case, the phase shift is also compensated, but not in a linear fashion.) Fig. 10-152A shows a typical series-peaked high-frequency compensated amplifier circuit. In this circuit, the peaking coil L_S is inserted in such a way that it isolates the shunting capacitances related to both tubes. *C1* represents those shunting capacitances relative to tube *V1*, and *C2* represents those shunting capacitances relative to tube *V2*. The combination of *C1*, L_S, and *C2* represents a filter network which has the characteristics of a low-pass filter. That is why series-peaking compensation is sometimes known as filter coupling.

Variations in the circuit of Fig. 10-152A are sometimes made in order to isolate the shunting capacitances at some specific ratio. For instance, in Fig. 10-152B, the series-peaking coil L_S is shifted to the plate of the load resistor because in some circuits the shunt capacitances are more evenly distributed under these circumstances. In Fig. 10-152C the peaking coil is placed on the grid side of the coupling capacitor C_b, so that the shunting capacitance between the capacitor C_b and ground will be added to *C1* and not to *C2* as in circuit (A). In all of the above cases, the series-peaking coil is inserted in such a way that the shunting capacitances are divided on either side of the coil with the ratio of *C1* to *C2* that is most beneficial to the circuit. For most practical purposes, and in our discussion, *C2* is equal to twice *C1*. In practice, the load resistor R_b is usually placed on the side of the lower shunting capacitance, namely *C1*. However, this is not always done, since the capacitances may be so nearly equal that it makes little difference. Since the total shunting capacitances are isolated by the coil L_S, the value of the load resistor R_b in most cases depends upon *C1*, whereas the value of the load resistor R_b in shunt peaking depended upon the total capacitance C_t. Therefore, the lower the value of *C1*, the higher the value R_b can be, and the greater the gain, since the gain depends upon the size of R_b. It then follows, from the above discussion, that both the values of *C1* and *C2* have to be estimated quite closely for the proper design and use of the plate-load resistor R_b and the series-peaking coil L_S.

The addition of the series-peaking coil helps the high-frequency response in the following way (refer to the circuit of Fig. 10-152A). The voltage output from tube *V1* is impressed across the plate-load resistor R_b. Assuming no compensation (L_S removed), then this voltage across R_b would be attenuated at the high frequencies due to the effect of the total shunting capacitance. With peaking coil L_S inserted, the total shunting capacitances are divided, with the smaller amount going to *C1*. The previous loss, which was due

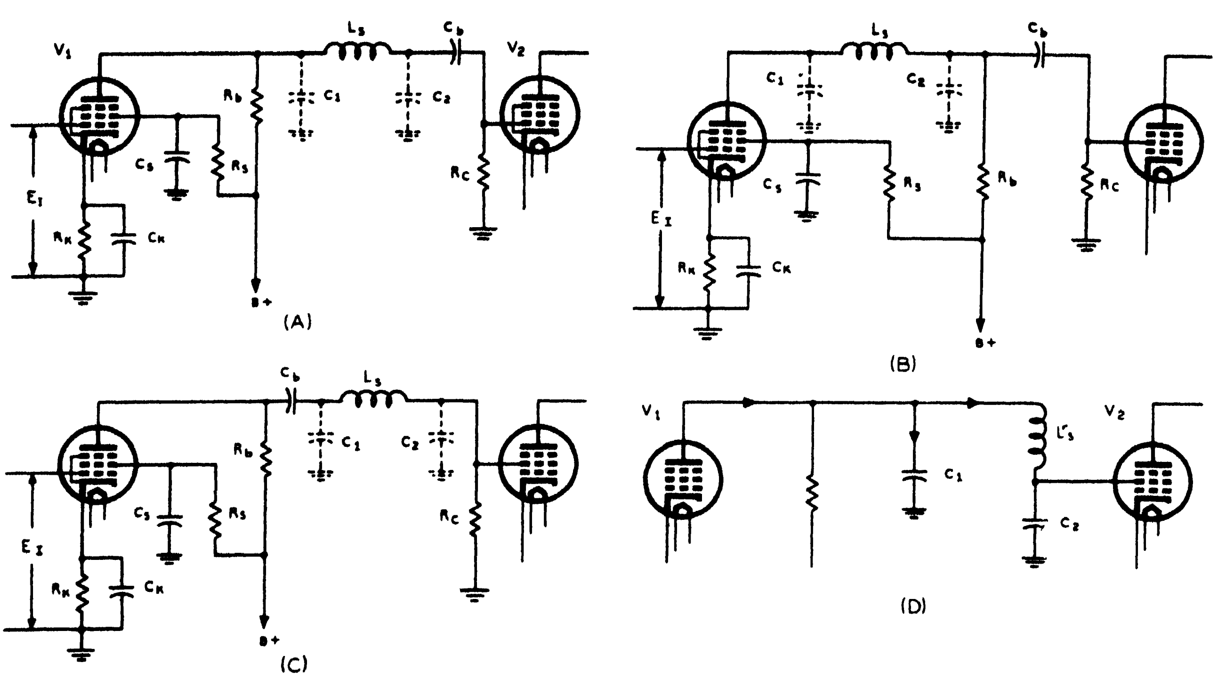


Fig. 10-152.—A typical series-peaking high-frequency compensated video amplifier circuit is shown in (A) and a simplification of this circuit in (D). The circuits in (B) and (C) are variations of (A), these being sometimes used to isolate the shunting capacitances at some specific ratio.

to the effect of the total shunting capacitances, is now greatly reduced by the effect of *C1* alone. This output voltage from tube *V1* that is now developed across R_b and *C1* in parallel is impressed on the voltage-dividing network of L_S and *C2*. C_b presents a short circuit at high frequencies and the value of the grid-leak resistor R_c is so large compared to the reactance of L_S and *C2* that both C_b and R_c can be considered as having negligible effect on the voltage dividing network L_S and *C2*. They are, consequently, neglected in the circuit analysis at high frequencies, even though the shunting capacitance across tube *V1* is reduced from the total shunting capacitance C_t. The problem now is to offset the high-frequency attenuation caused by *C1*. This can be done by offering an impedance to the output voltage of tube *V1* at high frequencies that is considerably lower than that represented by the R_bC1 combination, so that most of this signal output voltage will follow the path of low impedance and very little will be shunted to ground through *C1*.

The question that now arises is how this effect can be attained. Refer to Fig. 10-152D, which is a simplified circuit of Fig. 10-152A; at some frequency, call it f_o, we notice that L_S and *C2* will be series resonant. Therefore, at this frequency, the impedance offered by the circuit of L_SC2 to the output of tube *V1* will be a minimum, and *most* of the output signal voltage from tube *V1* will follow the path of minimum impedance offered to the tube, namely L_SC2, and very little will be shunted to ground by the higher impedance of *C1*. The signal current flowing from tube *V1* will then flow through the voltage dividing network of L_S and *C2*, and the maximum amount of voltage (at high frequencies) will be impressed on the grid of tube *V2* by the voltage drop across capacitor *C2*. It is, therefore, seen how capacitor *C2* offsets the drop in gain at high frequencies by the characteristics of its resonant effects, at these frequencies, with the series-peaking inductance L_S. In actual practice, the ratio of *C2* to *C1* is set approximately equal to two, as stated before. The values of *C1* can be measured fairly accurately, and so can the total capacitance C_t ($C_t = C1 + C2$), and from knowing *C1* and C_t, *C2* can be found. Circuit components may have to be shifted around in order to find the approximate ratio.

In working with design formulas for the series-peaking inductance and also for the load resistance, the operation of the series-peaking circuit of Fig. 10-152A may be better understood by discussing typical response characteristics at the high end of a curve — such as those shown in Fig. 10-153.

In this figure, the gain is only relative and its maximum value is taken as unity at the frequencies at the midpoint of the pass-band.

Curve *1* represents the actual curve of the amplifier without any compensation. It is readily seen how this curve falls off at the high frequencies. The horizontal dashed curve *3* is the 3-db curve where the gain drops

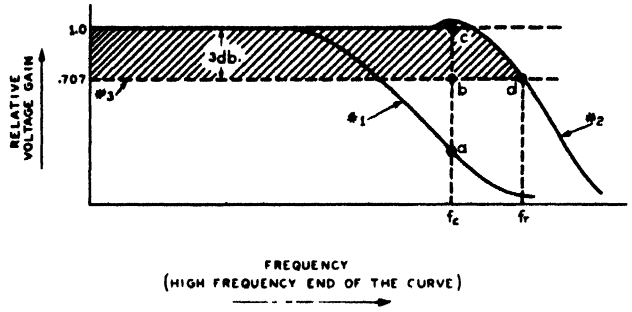


Fig. 10-153. — Typical response characteristic curves, wherein the voltage gain is relative, its maximum value being taken as unity at the intermediate frequencies.

to 70.7 per cent of its maximum value. (Keep in mind that this 3-db value is the point of minimum gain allowable for most proper designs.) Assume that we want to have a flat frequency response characteristic to a correction frequency f_c of 1 Mc. Then, from curve *1*, we note that, at this frequency, the gain is at point *a* and, therefore, undesirable. We want to bring up point *a* to somewhere in the shaded area. Ideally, we want to bring up point *a* in such a manner that curve *1* is raised at the high frequencies so it has a relative voltage gain of unity. In other words, even though upon correction, point *a* may advance anywhere into the shaded area (the area of allowable gain) between points *b* and *c*, it is preferable to approach point *c*. By actual experimental determination, it has been found (when $C2 = 2C1$) that this correction is ideally approached when the elements L_S and $C2$ are made to resonate at the correction frequency f_o. Under such conditions, curve *1* is shifted to curve *2*. From this curve, it is evident that at the correction frequency f_c, there is a peak rising *slightly* (which is due to series resonance between L_S and $C2$) above the level of unity gain; but, in general, the shape of the curve approaches the ideal case very closely.

The formulas for the design of the series-peaking coil L_S and the plate-load resistance R_b depend primarily on the value of the frequency desired for correction. Calling this frequency, as we have before, the correction frequency f_o, we know now that L_SC2 should be made to resonate at this frequency, whereupon we have

$$f_o = \frac{1}{2\pi\sqrt{L_SC2}} \qquad (10\text{-}25)$$

where the shunting capacitance $C2$ is known by measurement as discussed beforehand or by close approximations. Squaring both sides of the equation above and solving for L_S, we get

$$L_S = \frac{1}{(2\pi f_c)^2 C2}\,. \qquad (10\text{-}26)$$

Knowing that $C2 = 2C1$ and substituting this in equation (10-26), we get

$$L_S = \frac{1}{(2\pi f_o)^2(2C1)}\,. \qquad (10\text{-}27)$$

From the above equations, we can readily solve for the series-peaking inductance in terms of either shunting capacitance $C1$ or $C2$. We can rearrange equation (10-27) in the following manner

$$(2\pi f_o)L_S = \tfrac{1}{2}\cdot\frac{1}{(2\pi f_c)C1} \qquad (10\text{-}28)$$

which states that the inductive reactance of L_S at the frequency of correction is equal to ½ the capacitive reactance of $C1$ at the same frequency.

If we rearrange equation (10-27) in another way by solving the equation for the value $\sqrt{2}f_c$, we find the following.

Taking the square root of both sides of equation (10-27) we have

$$\sqrt{L_S} = \frac{1}{2\pi f_c\sqrt{2C1}} \qquad (10\text{-}29)$$

and solving for $\sqrt{2}f_c$, we get

$$\sqrt{2}f_c = \frac{1}{2\pi\sqrt{L_SC1}}\,. \qquad (10\text{-}30)$$

Referring to equation (10-30), it will be noticed that the right-hand side of the expression is itself representative of a resonant frequency. In other words, a frequency exists at which the series-peaking coil L_S is in resonance with the lower-valued shunting capacitance $C1$. The resonant frequency of L_SC1 is then equal to $\sqrt{2}$ times the frequency of correction. That is to say, if we let f_r equal the resonant frequency of L_S and $C1$, then

$$f_r = \sqrt{2}f_o. \qquad (10\text{-}31)$$

This means that there exists for L_S and $C1$ a resonant frequency f_r that is equal to $\sqrt{2}$ times the frequency of correction.

Knowing that the total shunting capacitance C_t is equal to the sum of $C1$ and $C2$ and that $C1 = C2/2$, then substituting for $C1$,

$$C_t = \frac{C2}{2} + C2 = \frac{3}{2}C2$$

and, solving this equation for *C2*, we get

$$C2 = \frac{2}{3}C_t$$

and, substituting this equation in equation (10-26), we get

$$L_s = \frac{3}{(2\pi f_o)^2(2C_t)} \;. \qquad (10\text{-}32)$$

Rearranging, we get

$$2\pi f_c L_s = \frac{3}{2} \cdot \frac{1}{2\pi f_c C_t}$$

or

$$X_{Ls} = \frac{3}{2}X_{Ct} = 1.5X_{Ct} \qquad (10\text{-}33)$$

which states that the inductive reactance of the series-peaking coil at the frequency of correction is equal to $\frac{3}{2}$ the reactance of the total shunting capacitance C_t at the frequency of correction.

In determining the value for the inductances of the peaking coil we can utilize any one of equations (10-25) through (10-33), according to the parameters that are known. These equations make use of the fact that in order to solve for L_s, all that has to be known are any two of the following elements that fit into the *particular formula* used

C_t = total shunting capacitance in the circuit $(C1 + C2)$

$C1$ = shunting capacitances around the output of the first tube

$C2$ = shunting capacitances around the input of the second tube

f_c = frequency of correction (or that frequency where L_s and $C2$ will be in series resonance)

f_r = that frequency which is $\sqrt{2}$ times the frequency of correction (or that frequency where L_s and $C1$ can be considered to be in resonance).

Now that we have decided on numerous ways for determining L_s, we are ready for the proper determination of the plate-loading resistor R_b in these series-peaking circuits.

From previous topics, we have determined that the plate-load resistance R_b should be equal to its shunting reactance for proper design purposes. In shunt peaking, this reactance was the total shunting reactance X_{Ct} in the circuit.

Therefore, for design purposes, R_b was set equal to X_{Ct} at the frequency of correction, which was the minimum acceptable 3-db frequency. One should understand that these assumptions are not just guesses but are based upon actual experimental data taken on such amplifier circuits. When the response curve of Fig. 10-153 was plotted, it was found by experimentation that the plate-load resistor R_b was approximately equal to *one-half* the reactance of the shunting capacitor *C1* at the frequency of correction. Accordingly, the basic design formula for the load resistor R_b for series peaking is

$$R_b = \tfrac{1}{2}X_{C1} = \frac{1}{2(2\pi f_o C1)} \;. \qquad (10\text{-}34)$$

Since it is easier to measure the total shunting capacitances C_t and *C1* alone, we shall rearrange equation (10-34) in terms of C_t instead of *C1* by substituting

$$C1 = \frac{C_t}{3} \text{ (since } C2 = 2C1\text{)}.$$

Hence

$$R_b = \frac{3}{2} \cdot \frac{1}{2\pi f_c C_t} = \frac{3}{2}X_{Ct} = 1.5X_{Ct} \qquad (10\text{-}35)$$

states that for proper series-peaking compensation, the plate-load resistor R_b is made equal to one and one-half times the total shunting reactance X_{Ct} at the frequency of correction (when $C2 = 2C1$).

We also can combine the relation between the equations (10-28) and (10-34) in such a manner that L_s and R_b are in the same equations. Thus

$$X_{Ls} = R_b \qquad (10\text{-}36)$$

and substituting $2\pi f_o L_s$ for X_{Ls} we get

$$2\pi f_o L_s = R_b \qquad (10\text{-}37)$$

where

$$L_s = \frac{R_b}{2\pi f_o} \,. \qquad (10\text{-}38)$$

Solving equation (10-35) for $2\pi f_o$ we get

$$2\pi f_o = \frac{3}{2C_t R_b} \qquad (10\text{-}39)$$

and substituting this equation in equation (10-38) we get

$$L_s = \frac{R_b}{3/(2C_t R_b)} = 0.67C_t(R_b)^2. \qquad (10\text{-}40)$$

From the above equations, it is seen that we can solve for the plate-load resistance R_b by knowing the frequency of correction f_o and any one of the quantities $C1$, C_t, or L_S.

Some fundamental properties of the series-peaking circuit are that it has 50 per cent more gain than a shunt-peaking circuit [compare equations (10-20) and (10-35)] where both circuits have the same values of f_o and C_t, and that the phase shift response in series peaking is more linear than in shunt peaking. Thus it is seen that series peaking betters the gain and phase characteristics of the amplifier at the high-frequency end of the curve as compared to shunt peaking.

Shunt-Series Peaking

Although series peaking results in better response and more linear phase shift than shunt peaking, a combination of these types gives a better response than either one alone. Fig. 10-154 illustrates a typical circuit combining the effect of series and shunt peaking with the shunting capacitances *C1* and *C2* shown. With this type of compensation, there is a sharper cutoff at the correction frequency than in the series case alone. Much greater gain is obtained than when shunt or series peaking alone is used. It has been found that combination peaking gives as much as 80 per cent more gain than a shunt-peaking system alone. This means that the total effective load impedance is also 80 per cent greater than that for shunt peaking.

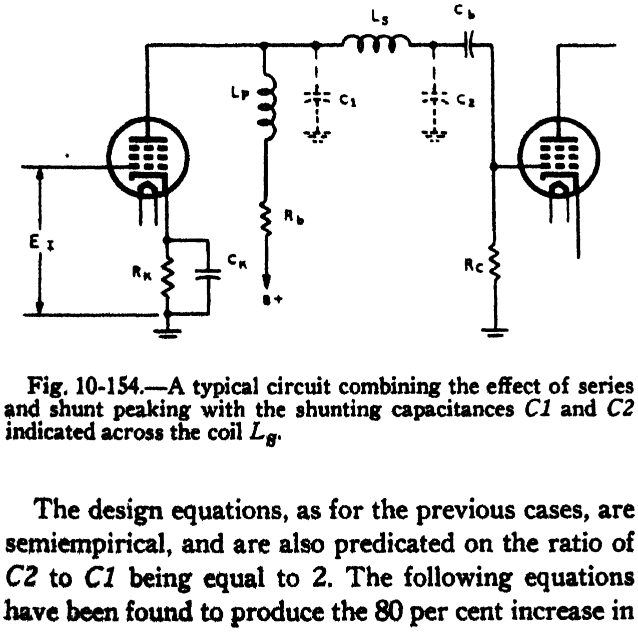


Fig. 10-154.—A typical circuit combining the effect of series and shunt peaking with the shunting capacitances *C1* and *C2* indicated across the coil L_S.

The design equations, as for the previous cases, are semiempirical, and are also predicated on the ratio of *C2* to *C1* being equal to 2. The following equations have been found to produce the 80 per cent increase in gain.[4]

Plate-load resistor

$$R_b = 1.8X_{Ct}. \qquad (10\text{-}41)$$

Shunt-peaking coil

$$L_p = 0.12C_t(R_b)^2. \qquad (10\text{-}42)$$

Series-peaking coil

$$L_S = 0.52C_t(R_b)^2. \qquad (10\text{-}43)$$

The Compensated Amplifier

So far, we have discussed the separate types of frequency compensation (high and low) which may be applied separately to any system desired. But in oscilloscope amplifiers, as previously pointed out, the desire is to have as flat a response as possible from the lowest audio frequency encountered to the highest frequency needed. It is evident that to obtain these qualities in a single circuit, the low-frequency and high-frequency compensating networks have to be combined. Fig. 10-155 illustrates a number of different amplifiers in which both low- and high-frequency compensations are employed. An interesting feature of these circuits is the additional capacitors used.

In all the circuits, it will be noted that the capacitors C'_k, C'_s, and C''_b are shunted across the capacitors C_k, C_s, and C'_b that were originally used in the low-frequency compensated case (see Fig. 10-148). One of the fundamental actions of capacitors C_k, C_s, and C'_b is to bypass low-frequency components to ground and, therefore, to prevent these frequencies from getting into certain parts of the circuit where they are not desired. Feedback is prevented by such bypassing. The values of these capacitors C_k, C_s, and C'_b are hence made large so that their reactance to low frequencies will be small. That is fine for low frequencies, but when the low-frequency compensated network is combined with the high-frequency compensated network, a separate path to bypass the high frequencies must be provided to prevent feedback. Two separate bypassing paths are thus needed, one for low frequencies and the other for high. We already have the low-frequency bypass capacitors in the circuit, and since bypassing of the high frequencies is necessary, then all that has to be done is to shunt capacitors C_k, C_s, and C'_b with the other capacitors C'_k, C'_s, and C''_b, respectively. Since these latter three capacitors are needed to bypass the high frequencies, they are much smaller than the former ones. Some typical values employed in practice are as follows:

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[4] Seeley, S. W. and Kimball, C. N., "Analysis and design of video amplifiers," *RCA Rev.*, vol. 3, pp. 290-308, January, 1939.

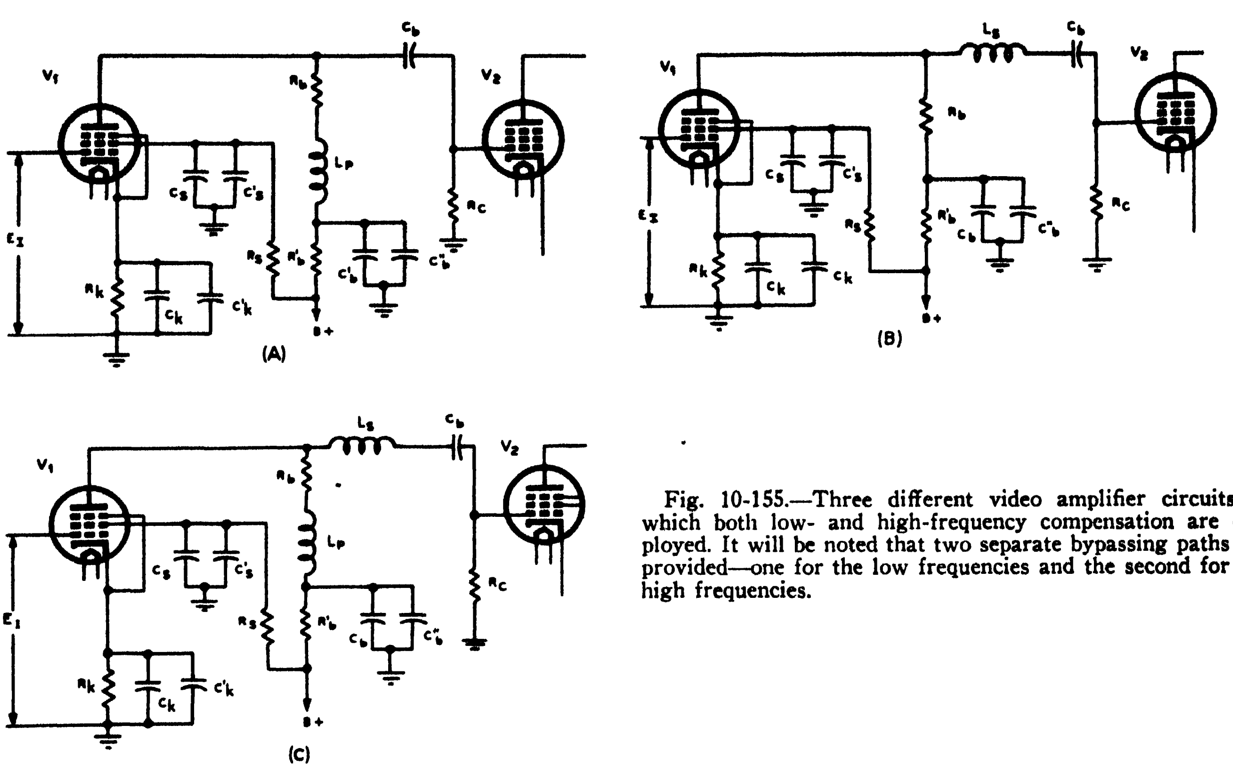

Fig. 10-155.—Three different video amplifier circuits in which both low- and high-frequency compensation are employed. It will be noted that two separate bypassing paths are provided—one for the low frequencies and the second for the high frequencies.

$C_k = 100\ \mu f$ $C_s = 16\ \mu f$ $C'_b = 30\ \mu f$

$C'_k = 0.1\ \mu f$ $C'_s = 0.01\ \mu f$ $C''_b = 0.1\ \mu f$

In visualizing these capacitors, one must remember that their size depends not only on the capacitance value but also on the voltage rating. For example, the 100-μf cathode bypass capacitor need only have a small value of breakdown voltage depending on the voltage drop across the cathode resistor R_k.

Assuming the bias on the tube to be −10 volts, then the drop across the bias resistor R_k is 10 volts, and for a factor of safety, the cathode bypass capacitor C_k (and likewise C'_k) usually has a voltage rating of 25 or 50 volts. Such a capacitor of 100 μf has to be of the electrolytic type and is not very large.

It is very interesting to note the similarity between very wide-band oscilloscope amplifiers and the video amplifiers utilized in television receivers. The difference between an oscilloscope amplifier with a top frequency of, say, 6 to 10 Mc and a video amplifier of the same range is insignificant. The requirements of both systems are almost the same; both must have linear phase shift characteristics and both require frequency compensations so as to afford the necessary response at the low and high ends of the band. If there is a difference, it may be found in the need for response at a lower frequency in the oscilloscope amplifier than in the video amplifier, but the means of accomplishing these conditions is virtually the same. The constants of the elements may not be the same, but the theory is. These comments are made to indicate that texts which describe the process of frequency compensation in video amplifiers have their application to the same end in oscilloscope amplifiers.

As an example of this close relationship, Fig. 10-156 illustrates a system of high- and low-frequency compensation which is the equivalent of the circuits employed in video amplifiers, differing only in the specific constants. The low-frequency compensation is accomplished by resistor *R* and capacitor *C* in the plate circuit. This corresponds to R'_b and C'_b in the previous discussion of low-frequency compensation. The plate-load resistor *R1* corresponds to R_b, the coupling capacitor *C1* corresponds to C_b, and the grid-load resistor R_o is the equivalent of *R2* in parallel with *R3*. If the compensating circuit were not used, the gain at the low frequencies would fall off due to the effect of the *C1-R2-R3* network. The reactance of capacitor *C1* goes up as the frequency goes down, causing less of the output voltage of the first stage to appear across the grid load of the second stage. To correct this, *R* and *C* are chosen so that the impedance of this combination increases with decreasing frequencies in the proper ratio to compensate for the falling off due to the coupling circuit feeding the next stage. This compensation occurs because a higher load impedance in the plate circuit results in a greater output. The values of resistance and capacitance are also chosen to make the phase shift

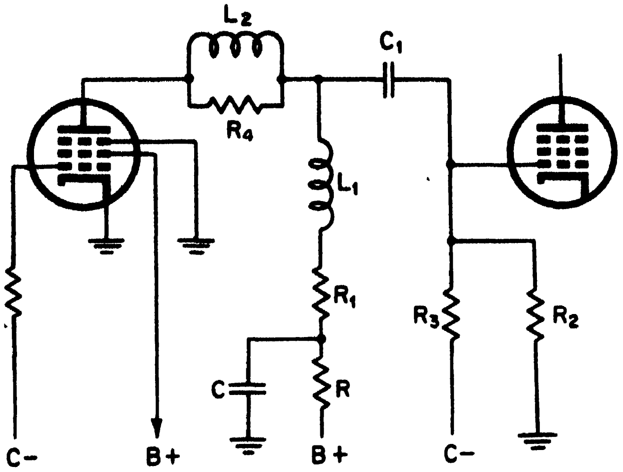

Fig. 10-156.—Typical wide-band oscilloscope amplifier employing high- and low-frequency compensation; its similarity to a video amplifier is evident.

linear with respect to frequency. The value of R is limited by the drop in voltage across this resistor, which decreases the plate voltage.

The high-frequency compensating combination employs both series and shunt peaking. This is done so that the response at the high frequencies will be better than it would be if only shunt or series peaking were used. The shunt-peaking system used in Fig. 10-156 is similar to that shown in Fig. 10-150A. The peaking coil *L1* is equivalent to L_p in the previous discussion. This coil resonates with the shunt capacitance, which includes all the stray capacitance associated with the wiring of that circuit. It is difficult to estimate the specific value of this capacitance, as it varies quite a bit, although reference information relative to the tube is available, and the total is determined by experiment. (It is in this connection that lead dress is an important constructional detail.) In the actual design of a circuit such as this, experimental verification complements the measurements.

The series-peaking coil *L2* is the equivalent of the peaking coil L_S in the previous discussion of series peaking. The coil resonates with the shunt capacitances to produce a peak at the high end of the frequency range; this peak, together with the peak caused by the shunt-peaking coil, is sufficient to flatten the response up to the highest frequency for which the constants were selected. The resistor *R4* connected in parallel with *L2* serves to reduce the Q of that coil, or to damp the resonant circuit so as to cause the desired decay of the signal. This prevents possible oscillation or ringing in the system. The use of such a damping resistor is not a must — it is dependent upon the conditions in the circuit, or upon the height of the resonant peak. If it is too high and can cause ringing, the resistor is used. Examples of both are to be found in oscilloscope amplifiers.

It may appear from all that has been said about frequency compensation that all of it is accomplished in a single stage. Such is not the case; every stage in a compensated system makes use of some compensation, although the resonant peaks in each stage may not necessarily be the same. The compensation is determined by the bandwidth required and by the performance of the different stages relative to the degree of attenuation in each stage. In some instances, the design may be such that the resonant frequencies are staggered in the different stages, so that each makes a contribution at one frequency in the amount necessary to offset the loss in another. This tends to develop a reasonably flat response over a wide band.

If there is a critical point in the system, it is found in the output stage which feeds the cathode-ray tube, especially if the tube is of substantial screen size so that fairly long interconnecting leads are required. Such connections mean increased values of distributed capacitance, and while the preceding stages may afford the desired response curve, the increased limitation due to the higher capacitance in the output system demands the utmost in correction.

CHAPTER 11

SYNCHRONIZATION

In Chapter 9 we explained how the frequency of the sweep circuit had to be adjusted to that of the incoming signal or some multiple thereof in order to reproduce one or more cycles of the input signal. Nothing, however, was said about keeping the pattern stationary upon the screen of the cathode-ray tube. If after adjustment of the coarse and fine sweep-frequency controls, the sweep frequency remained *constant,* then the pattern would remain stationary. However, such is usually not the case. Let us analyze why this is so.

Nature of Pattern without Synchronization

The sawtooth waveform produced by relaxation oscillators is very rich in harmonics of the fundamental oscillator frequency; because of this, the oscillator is not very stable.[1] Therefore, the ability of a pattern to remain stationary is primarily dependent upon the stability of the sweep oscillator.

Assume that a sinusoidal pattern has been "stopped" on the screen by simple manual operation of the sweep-frequency controls, and after a period of time the frequency of the sweep begins to drift. What happens to the picture? When the pattern is stopped, it means that the beginning of the sawtooth trace will constantly repeat itself at the same point on the input signal. In other words, the phase relationship between the sawtooth sweep and the input signal is held constant. When the sweep frequency begins to drift, the phase relationship changes, and if the sawtooth sweep has a slow continuous drift there is a varying phase shift between the two signals. Due to this frequency drift, and hence phase shift, the start of the sweep trace will not occur at the same time for each cycle or group of cycles of input signal, but rather at different successive points along the cycle. The result of this action is that the pattern will appear to be moving across the screen from right to left, or left to right, as the case may be.

The rate of travel across the screen is dependent upon how great the phase shift becomes from its original position. When this phase shift is slight, the pattern moves slowly across the screen, and when the phase shift is great the pattern moves very rapidly. In order to illustrate what the patterns would appear like, a number of oscillograms were photographed under different conditions of phase shift, with the input signal being a sine wave. These photos appear in Fig. 11-1. All pictures were exposed the same amount of time, approximately one second, so that the "thickness" of the resultant wave on the photograph indicates the relative amount of phase shift.

The oscillograms at (A) and (B) indicate instances where the fine sweep-frequency control of the oscilloscope was varied a small amount from the position where it just about held the pattern stationary. The phase shift for that at (B) is slightly greater than that at (A). The patterns at (C) and (D) indicate instances where the phase shift is great and the waves are traveling across the screen very rapidly. This rate of travel compared to the shutter speed of the camera is so fast that the reproduced pictures at (C) and (D) appear as solid patterns, often referred to as "smear" patterns.

From the above analysis we see that, in order to keep the pattern on the screen stationary, the sweep oscillator must be stable. In the sweep circuits of oscilloscopes and television receivers, the stability of these circuits is brought about by a process called *synchronization.* Generally speaking, synchronization means the injection of a special signal or *control voltage* into the oscillator circuit to stabilize its frequency of operation. When the frequency is synchronized, it is said that the oscillator "locks-in" with the control signal. When used with oscilloscopes, synchronization brings about the *constant timing* of the input signal with that of the sweep signal so that their cycles of operation coincide with each other.

The application of the synchronizing signal, usually just referred to as *sync signal,* varies somewhat with the type of sweep circuit employed. Although synchronization is employed in many different types of pulse-forming circuits, in this chapter we will only consider

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[1] Groszkowski, J., "The interdependence of frequency variation and harmonic content, and the problem of constant-frequency oscillators," *Proc. I.R.E.*, vol. 21, pp. 958-981, July 1933.

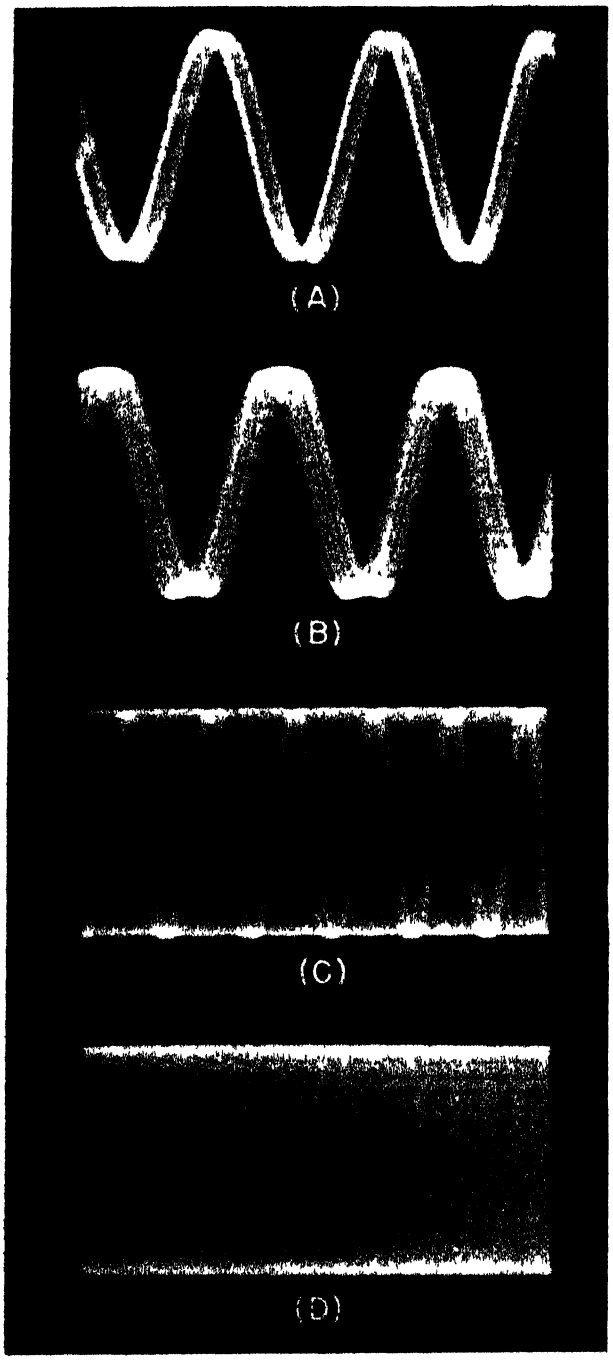

Fig. 11-1.—Changing phase conditions between sawtooth and input sine wave causing unsynchronized patterns. The phase difference increases from (A) to (D).

synchronization of sawtooth oscillators of the gaseous relaxation type, multivibrators, and blocking oscillators. With sawtooth oscillators, the sync signal is applied to the circuit in such a manner that it will initiate the trace or retrace of the sawtooth wave. In most cases, as in the circuits included in Chapter 9, the sync signal initiates the retrace. In other words, it starts the current flowing in the discharge tube and hence starts the discharge of the capacitor.

SYNCHRONIZATION OF THYRATRON SAWTOOTH OSCILLATOR

Synchronization in oscilloscopes is brought about by taking some of the signal to be observed and feeding it to the sweep circuit of the unit. This signal then becomes the sync signal and is usually fed to a grid circuit of the sweep oscillator. In order to completely understand the fundamental operation of synchronization, we will theoretically analyze this phenomenon for the simple circuit of a thyratron sawtooth oscillator and then prove the theory by experimentation. The sweep oscillator will only synchronize or "lock-in" with the incoming signal when the frequency of this signal is equal to, or is a multiple of, the *synchronized* sawtooth sweep frequency.

Relationship between Grid Voltage and Ionization Potential

In Chapter 9 we proved by way of experimentation how the grid bias on the thyratron controlled the ionization potential of the tube (see Table I in Chapter 9). If the grid bias was made more negative, the ionization potential was increased, and when made less negative the ionization potential decreased. The variation of ionization potential with grid bias is a constant one. In other words, the grid-voltage versus plate-voltage characteristic of the thyratron is a linear one. Thus if a grid-bias change from −8 volts to −12 volts produces a change in ionization potential from 70 to 110 volts, an increase of 40 volts, then an additional increase in the grid bias from −12 volts to −16 volts should produce a new ionization potential of 110 volts plus 40 volts, or 150 volts. In a case such as this, we see that every 4-volt change in grid bias is accompanied by a 40-volt change in ionization potential, which means that the grid controls the ionization potential at a ratio of one to ten.

We can express this voltage relationship for any thyratron mathematically. If E_i represents the ionization potential of the tube and E_g the negative grid bias, we have

$$k = \frac{E_i}{E_g} \qquad (11\text{-}1)$$

where k is the ratio of increase in ionization potential to change in grid bias and is always greater than unity. Since k, often called the *grid-control ratio,* is constant, it represents the slope of the linear E_i-E_g characteristic of the tube in question.

The value of *k* for any one tube tells us how much the amplitude of the sawtooth voltage will change with a variation in grid bias. Since a change in ionization potential also produces a change in the frequency of the sawtooth output signal, it must be remembered that any variation in grid bias will, therefore, change the frequency of the sawtooth (see Table I in Chapter 9).

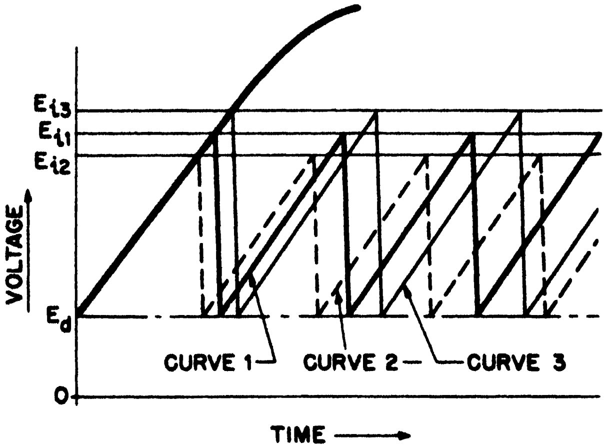

Fig. 11-2.—A change in grid bias and hence ionization potential causes a change in the frequency and amplitude of the sawtooth output of a thyratron.

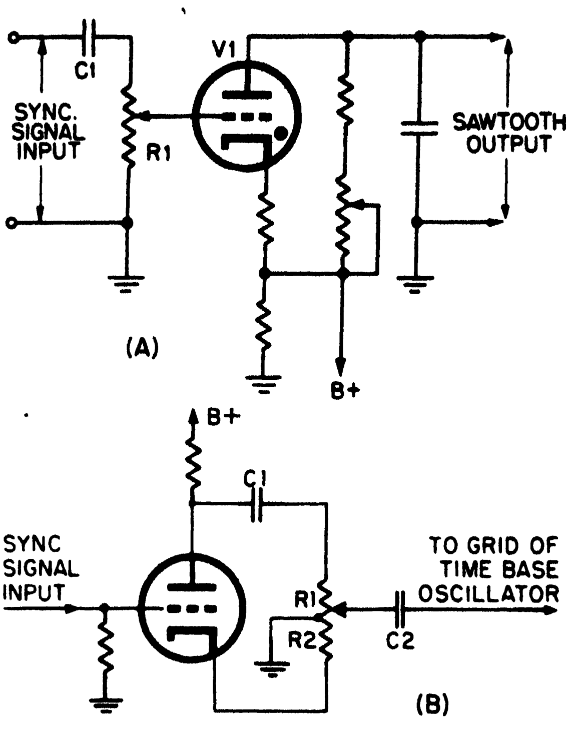

Fig. 11-3.—Application of a sync signal to a typical thyratron circuit is shown at (A); the circuit of (B) allows for choice of sync signal of any polarity.

The drawing of Fig. 11-2 illustrates the effect that three different values of grid bias have upon the ionization potential of the tube, and also upon the amplitude and frequency of the sawtooth output signal. Assume that the initial operating bias of the tube is such that the ionization potential is represented by voltage E_{t1} and results in sawtooth curve *1*. A negative bias decrease is indicated by the decreased ionization potential E_{t2} which results in sawtooth waveform *2*. An increase in the negative bias increases the ionization potential as indicated by voltage E_{t3}. Curve *3* results from this increase in ionization potential. Examination of curve *2* indicates a reduction in amplitude and an increase in frequency over curve *1*, and curve *3* indicates an increase in amplitude and a reduction in frequency over curve *1*. A change in bias does not change the deionization potential of the tube. This is indicated in the drawing of Fig. 11-2, where E_d represents the deionization potential of the tube.

Now that we have seen how the bias voltage on the thyratron affects the ionization potential of the tube and hence the frequency and amplitude of the sawtooth, let us put this knowledge to use in analyzing the process of synchronization as applied to this tube. A typical circuit of a thyratron sawtooth oscillator showing the application of a sync signal to its grid appears in Fig. 11-3A. This circuit is the same as that shown in Chapter 9 except that capacitor *C1* and resistor *R1* are employed to couple the sync signal to the grid of the tube. Resistor *R1* is made variable so that the strength of the sync-signal input can be controlled.

Certain types of sweep circuits require a specific polarity of sync signal for proper operation. The circuit of Fig. 11-3A does not allow for such polarity changes. To provide for a single control that will enable choice of sync signals of either polarity, many oscilloscopes utilize a circuit similar to that of Fig. 11-3B. The sync signal is applied to the grid of the tube which is connected in the form of a phase-splitter circuit. Between plate and cathode is a potentiometer which has a grounded center tap; that is, *R1* = *R2*. The variable arm of the potentiometer is capacitor-coupled to the grid of the time-base oscillator. (In some cases it may first be coupled to a sync amplifier and then to the time-base oscillator.)

By having the potentiometer center-tapped to ground, equal sync signal voltages appear across *R1* and *R2*. Regardless of the polarity of the input signal, the sync signal voltages across *R1* and *R2* will be of opposite polarity. That across *R1* will be 180° out of phase with the input sync signal and that across *R2* will be in phase with the input sync signal. By the potentiometer arm being capable of variation across either *R1* or *R2*, sync signals of varying amplitude and either positive or negative polarity are, therefore, available from one control. *C1* and *C2* also function as d-c blocking capacitors as well as a-c coupling units.

Since part of the input signal being observed is utilized as the sync voltage, it should, therefore, be remembered that *the sync signal can have any shape or form and still perform its required service.* However, for the sake of illustration in the following discussion and drawings, we will employ a sine wave as the signal to be reproduced and hence as the sync signal applied to the time-base oscillator.

Sine-Wave Synchronization

Let us assume that a sine-wave signal is to be observed and that some of this voltage is being used as the sync signal for the sweep circuit. This means that the bias on the thyratron tube will be varying in accordance with the applied sync signal. If the fixed bias on the thyratron is −5 volts and if the sync signal has a peak-to-peak voltage of 2 volts, then the bias will be varying between −4 and −6 volts. Since the ionization potential is determined by the bias on the tube, it, too, will be varying sinusoidally. However, the sinusoidal variation of the ionization potential, or plate voltage, is 180° out of phase with that on the grid, because as the grid goes more negative, the ionization potential is increased, (i.e., goes more positive), and as the grid bias becomes less negative (i.e., it increases in the positive direction), the ionization potential decreases.

Refer to Fig. 11-4. The straight line curve in this drawing is a typical E_p-E_g characteristic of a thyratron tube.[2] The sine wave at the lower part of the drawing indicates the sync signal input which is seen to be superimposed over the fixed grid bias of the tube. The ionization potential of the tube is represented by the sine wave at the right side of the drawing. Note that this curve is 180° out of phase with the input signal.

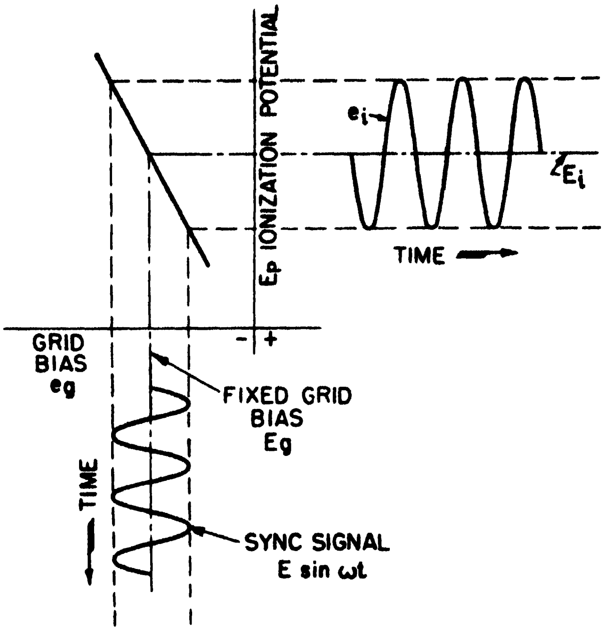

Fig. 11-4.—A typical E_p-E_g characteristic of a thyratron tube, indicating how the ionization potential changes with a sinusoidal change in grid bias.

If the fixed negative bias on the grid of the thyratron is represented by $-E_g$ and the applied sync signal represented by $E \sin \omega t$, then the varying bias voltage e_g can be represented as

$$e_g = -E_g + E \sin \omega t \qquad (11\text{-}2)$$

where E is the peak voltage of the sync signal. Since the ionization potential is 180° out of phase with the signal at the grid, then the varying ionization potential of the tube e_i can be written as

$$e_i = k\ (E_g - E \sin \omega t). \qquad (11\text{-}3)$$

By substituting equation 11-1 for E_g in equation 11-3, this latter equation becomes

$$e_i = E_i - kE \sin \omega t \qquad (11\text{-}4)$$

where E_i is the fixed ionization potential of the tube which is determined solely by the fixed bias on the tube (i.e., before application of the sync signal).

In the following analysis of synchronization, the sawtooth sweep signal will be assumed to have a constant rate of charge (i.e., linear rise time) and negligible flyback time. This assumption is necessary in order to make the discussion easier to understand. The frequency of the sawtooth sweep without synchronization will be referred to as the *natural* or *free-running frequency* of oscillation and with synchronization it will be referred to as the *synchronized sawtooth frequency.* The following is a list of notations that will be used throughout this chapter:

f_s = frequency of the applied sync signal
f_o = synchronized frequency of relaxation oscillation
f_n = natural or free-running frequency of relaxation oscillation
$n = f_s/f_o$
E_i = fixed ionization potential (without sync signal applied)
E_d = deionization potential

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[2] In this chapter the plate voltage E_p of the thyraton is more often referred to as the ionization potential and is designated E_i.

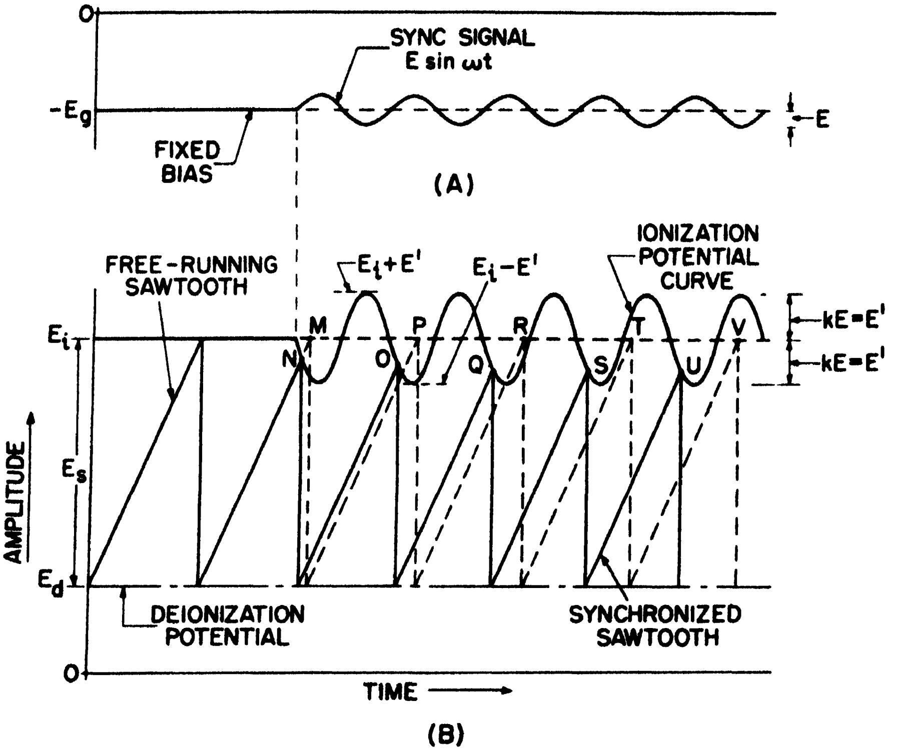

Fig. 11-5. — Graphical representation of the process of synchronization where the sync-signal input is sinusoidal.

$E_s = (E_i - E_d)$, the amplitude of the sawtooth wave before the application of the sync signal

$E' = kE$, the peak amplitude of the sinusoidal ionization potential (after application of the sync signal)

$m = E'/E_s$, ratio of the peak amplitude of sinusoidal ionization potential to the amplitude of the sawtooth signal before application of sync signal.

In the proper operation of the oscilloscope before the application of a sync signal, the sweep-frequency control is adjusted until the pattern on the screen is as stationary as possible. (If the pattern on the screen becomes stationary, then at that moment the natural frequency of relaxation oscillations is equal to some integral submultiple of the sync signal, that is, $f_0 = f_s/n$. The integral relationship depends on the number of cycles present upon the screen of the tube.) After this is done, the sync signal is then applied to the circuit to insure a locked-in effect between the input signal and the sawtooth sweep. Therefore, the frequency of the sync signal will differ by a *small amount* from some integral multiple of the normal frequency of relaxation oscillations, whether this integral be one, ten, etc.

The drawing of Fig. 11-5 basically illustrates this process of synchronization. Part (A) indicates the graph of bias voltage of the tube. The application of a sync signal is shown by the sine curve superimposed over the fixed negative bias $-E_g$. The sine wave in part (B) indicates the 180° out-of-phase alternating ionization potential of the thyratron tube which is seen to be superimposed over E_i. The E_i line is, therefore, said to represent the baseline of the alternating ionization potential. If E represents the peak voltage of applied sync signal, then kE is the peak voltage of the alternating ionization potential, where k is the grid-control ratio. For the sake of simplicity, we are making $kE = E'$, as indicated in the drawing. The maximum ionization potential will, therefore, be equal to $(E_i + E')$ and the minimum ionization potential equal to $E_i - E')$. The deionizing potential of the tube is constant and is represented by E_d; $(E_i - E_d)$, the amplitude of the sawtooth without synchronization, is indicated by E_s.

The sawtooth wave at the extreme left of part (B) represents the free-running frequency of oscillation. Without any sync signal applied, the tube constantly fires at E_i. In this drawing, the frequency of the sync signal is seen to be slightly greater than that of the free-running sawtooth sweep. With the application of a sync signal, the ionizing potential of the tube changes. The discharge of the sawtooth which would normally occur at point M along the constant E_i potential now occurs at point N, because the ionization potential is decreased at this latter point. After the discharge, the capacitor charges up again at the same rate as it did previously until the tube fires at point *O,* at which time the capacitor once again discharges.

If the sawtooth sweep continued to oscillate at its free-running frequency, the tube would fire successively at points *M, P, R, T,* and *V* along the fixed ionization potential of E_i as shown in the drawing. The sawtooth sweep shown dashed is a continuation of the free-running sweep frequency as it would appear without any sync signal applied. The sync signal, however, after point *N,* causes the tube to fire successively at points *O, Q, S,* and *U,* which are all reduced ionization potentials. At the very beginning of these new firing points, the reduced breakdown potential will be slightly lower than that preceding it, but after a number of cycles of applied sync signal, the tube will subsequently fire at the same reduced potential. This is indicated by the sawtooth waves at the right side of the drawing. When this occurs, the sawtooth signal is said to be locked-in with the applied sync signal and synchronization is complete. For the case under discussion, when synchronization occurs, the frequency and hence the period of the sync signal and that of the synchronized sawtooth sweep are equal to each other, that is, $f_s = f_o$. This can easily be checked from the drawing of Fig. 11-5.

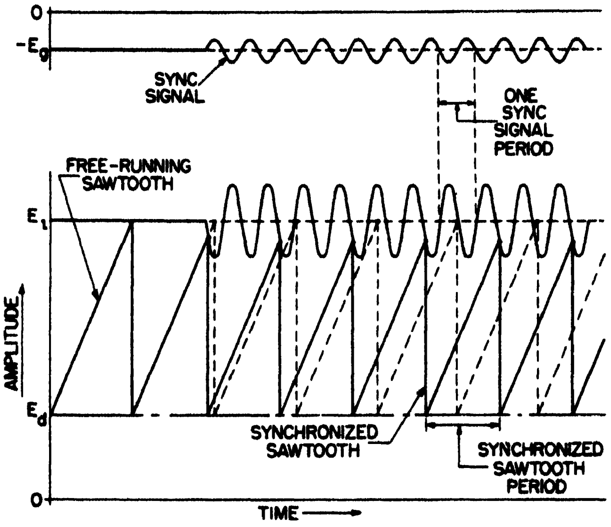

Fig. 11-6.—The frequency of the sync signal is slightly greater than twice the free-running sawtooth frequency in this graphical analysis of synchronization.

In order to indicate how synchronization can occur for the case where the frequency of the sync signal is slightly greater than twice the natural frequency of relaxation oscillations, refer to Fig. 11-6. The periods of the free-running sawtooth and also of the synchronized sawtooth are the same as those of Fig. 11-5. A measurement of the synchronized sawtooth period of Fig. 11-6 will readily indicate that it is equal to exactly twice that of the sync signal period, indicating that the frequency of the synchronized sawtooth signal is exactly equal to one-half that of the sync signal and that synchronization is complete; in other words, $f_s = 2f_o$ or $n = 2$.

Unstable Synchronization

Any small change in the operating frequency of a synchronized relaxation oscillator will cause the oscillator to be momentarily pulled out of synchronization. When this occurs, we can say that the situation is analogous to the beginning of Fig. 11-5 where the frequency of the sync signal and of the free-running sawtooth wave are not equal to each other. However, due to the presence of the sync signal, the frequency of the relaxation oscillator will be rapidly locked-in synchronization once again. In other words, the presence of the sync signal, once synchronization is obtained, assures the locking-in effect between the relaxation oscillator and the signal to be observed. This stability of synchronization will occur only when the tube fires along a *negative slope* of the alternating ionization potential curve.[3]

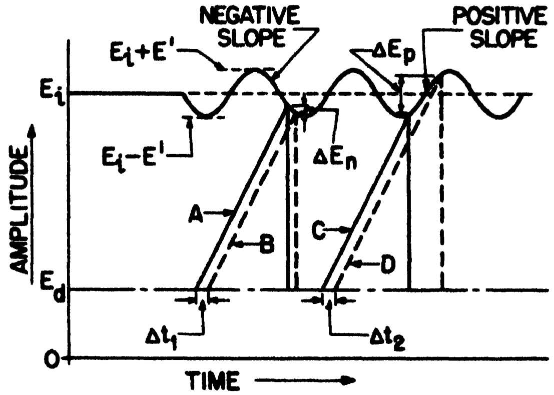

Fig. 11-7.—Unstable synchronization will occur when the sawtooth fires along the positive slope of the sinusoidally varying ionization potential.

Let us examine the drawing of Fig. 11-7. Those parts of the curve from $(E_i + E')$ to $(E_i - E')$ represent the negative slopes of the ionizing potential. They are characterized by the voltage going in the negative direction (i.e., less positive). Consequently, those parts of the curve from $(E_i - E')$ to $(E_i + E')$ represent the positive slope. Assume that sawtooth curve *A* is already synchronized and that it constantly fires at the same point along the negative slope of the ionizing curve. Assume further that the frequency of the relaxation oscillator drifts slightly, such that the new sawtooth wave is represented by dashed curve *B*. This frequency drift is indicated by the time change of Δt_1. Notice that this small frequency drift, in turn, causes only a small change in ionization potential which is indicated by ΔE_n. Due to this small change in ionizing potential, it is a simple matter for the sync signal, after a short period of time, to bring the oscillator back into synchronization.

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[3] MacLean, W. R., "The synchronization of oscilloscope sweep circuits," *Communications,* vol. 23, pp. 23-30, March, 1943.

Now let us look at a situation where the synchronized sawtooth fires along the positive slope of the ionizing potential. Curve *C* of Fig. 11-7 represents such a situation. For comparison purposes, the slope of curve *C* is made the same as curve *A*. Curve *C* is in synchronization, but is considered as being at an *unstable point of synchronization.* Let us again assume that the relaxation oscillator drifts in frequency so that the new sawtooth wave is indicated by dashed curve *D*, where the frequency drift is the same as that existing between curves *A* and *B*. Therefore, Δt_2, which represents the period change between curves *C* and *D*, is equal to Δt_1.

Notice the difference in firing potential between curves *C* and *D*. This difference, indicated by ΔE_p, is much greater than ΔE_n. The amplitude of sawtooth wave *D* is so much greater than that of curve *C* that once the relaxation oscillator drifts in frequency by the small amount indicated, it will be very difficult for the sawtooth to return to its original condition of equilibrium. The sawtooth wave has a tendency to seek a negative slope on the ionization potential curve as a point of stable equilibrium.

Limits of Synchronization[4]

Previously, we stated that the frequency of the sync signal usually differs by a small amount from the natural sawtooth frequency or some integral multiple of it. There are limits, however, as to how great this frequency difference can be and still produce effective synchronization. These limits must lie along the negative slope of the ionizing potential curve because of stability of synchronization. In other words, the synchronized sawtooth wave may have its striking potential lie anywhere along the decreasing slope from $(E_i + E')$ to $(E_i - E')$. Written in another manner, we can state that the ionizing potential, at which firing occurs in every relaxation cycle, can be defined by any value of voltage equal to $(E_i + E' \sin \theta)$ where θ may have any value from 90° to 270°.

From the above, we see that we have certain limiting conditions under which synchronization will be effective. In Fig. 11-8, we have indicated three individual synchronized sawtooth waves. In each case, the frequency of the synchronized sawtooth is equal to twice that of the sync signal; but you will notice that the firing point for each sawtooth occurs at different points along the ionization potential curve. Curves *A* and *C* indicate the possible upper and lower limits of firing voltage for stable synchronization. Any sawtooth wave initially falling in between these limits will also be synchronized. Curve *B* represents the intermediary sawtooth wave.

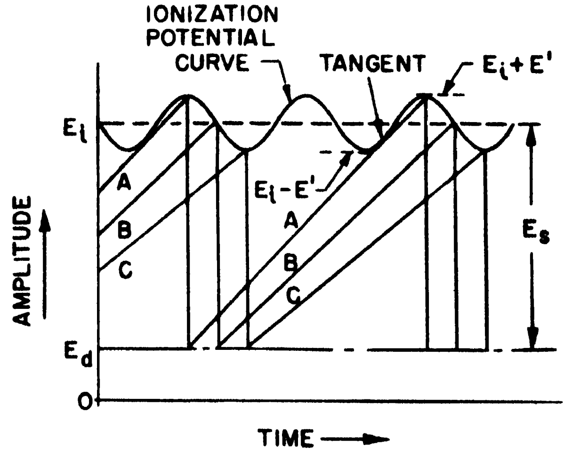

Fig. 11-8—Limiting conditions of stable synchronization where curves *A* and *C* represent the upper and lower limit of the natural frequency of relaxation oscillations.

Although the frequencies of synchronized sawtooth waves *A*, *B*, and *C* are all the same, the free-running frequency of each is different. Consequently, from the above analysis, we see that synchronized sawtooth curves *A* and *C* represent the *frequency limits of natural relaxation oscillations* that will still permit synchronization. Curve *B* in this drawing is shown to intersect the ionizing potential at E_i and this intermediary curve represents the case where the free-running frequency of relaxation oscillations f_n is equal to f_o, the synchronized sawtooth frequency. Therefore, for curve B, $f_n = f_o$.

If $f_n > f_o$, then the synchronized sawtooth will cause the breakdown of the thyratron anywhere between curves *B* and *C*, where curve *C* indicates the lower frequency limit of f_n. In like manner, if $f_n < f_o$, then the firing point can lie anywhere between curves *A* and *B*, where curve *A* indicates the maximum possible upper limit of f_n. The limiting frequency difference between f_n and f_o is, however, very small.

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[4] Builder, G., and Roberts, N.F., "The synchronisation of a simple relaxation oscillator," *A.W.A. Tech. Rev.* (Australia), vol. 4, no. 4, pp. 170-180, 1939; the material in this and some of the following sections is primarily based on this paper.

It is the amplitude of the sawtooth signal which is of more interest to us rather than the ionization potential E_i in the discussion of the frequencies of relaxation oscillations. This is so because the slope of the sawtooth trace helps define its frequency of operation. In this discussion, the amplitude of the sawtooth wave before synchronization is represented by E_s; therefore, from Fig. 11-5 we see that $E_s = (E_i - E_d)$. From this drawing, we see that the amplitude of the *synchronized* sawtooth e_s can be represented by

$$e_s = E_s + E' \sin \omega t. \qquad (11\text{-}5)$$

The polarity of the sign in front of $E' \sin \omega t$ depends upon the polarity of the applied sync signal. If the sync signal is a negative sine wave, then the sign will be positive as in equation (11-5). If the sync signal is an ordinary sine wave where the positive peak amplitude is at the 90° point of the cycle, then the sign will be negative [see equation (11-4)].

Consequently, it can be shown from the basic analysis relative to Fig. 11-8, that the minimum frequency of the free-running sawtooth wave is

$$\begin{aligned} f_{n\,(\min)} &= f_o\left(\frac{E_s - E'}{E_s}\right) \\ &= f_o\left(1 - \frac{E'}{E_s}\right) \\ &= f_o\ (1 - m). \qquad (11\text{-}6) \end{aligned}$$

In like manner, the maximum possible natural sawtooth frequency can be represented as

$$\begin{aligned} f_{n\,(\max)} &= f_o\left(\frac{E_s + E'}{E_s}\right) \\ &= f_o\left(1 + \frac{E'}{E_s}\right) \\ &= f_o\ (1 + m). \qquad (11\text{-}7) \end{aligned}$$

Equations (11-6) and (11-7) can be written in terms of the sync signal frequency f_s and also in terms of n, which is the frequency ratio of the sync signal to synchronized sawtooth signal. Thus

$$f_{n\,(\min)} = \frac{f_s}{n}\ (1 - m) \qquad (11\text{-}8)$$

$$f_{n\,(\max)} = \frac{f_s}{n}\ (1 + m). \qquad (11\text{-}9)$$

With $f_{n\,(\max)}$ represented by curve *A* of Fig. 11-8 and $f_{n\,(\min)}$ by curve *C*, then the frequency differences of $(f_{n\,(\max)} - f_o)$ and $(f_o - f_{n\,(\min)})$ are the same. This maximum value of natural relaxation oscillations is, however, only possible if the ratio of E'/E_s or m is small enough to allow the trace or charging part of the sawtooth to reach the maximum ionization potential of $(E_i + E')$ without having reached the ionization potential at an earlier instant. Since the ionization potential of E_i is fixed by the fixed d-c bias on the thyratron tube, then in order to keep the ratio m very small, the amplitude of the sync signal must be kept low.

Curve *A* of Fig. 11-8 best illustrates the situation. From this curve, it is obvious that any small increase in the amplitude of the ionization potential curve (which is brought about by an increase in the sync signal amplitude) and hence m, will cause curve *A* to strike in the vicinity of the negative peak of the previous half-cycle of ionization potential. In other words, the thyratron circuit capacitor representing curve *A* would discharge around the minimum ionization potential $(E_i - E')$ of the previous half-cycle.

From Fig. 11-8 we also see that sawtooth wave *A* is tangent to the ionization potential curve along the positive slopes of the sine wave. This tangent point determines what the maximum permissible free-running relaxation oscillations can be. If something happens to the relaxation oscillator so that the positive slope of curve *A* is decreased slightly, then curve *A*, still being tangent to the sine wave, will no longer strike at point $(E_i + E')$, but at some smaller value. Thus the maximum permissible free-running relaxation frequency will be less than before, and curve *A* will appear to pivot ever so slightly in the clockwise direction, with the pivot point being at the tangent between the two curves.

It can be shown by geometrical means that m cannot be greater than approximately $1/(4.3n - 1)$ in order to attain the maximum possible ionization potential of $(E_i + E')$. From this relation, it is evident that if n is a large number, that is, when the ratio of f_s/f_o is high, the value of m must be quite small to obtain the maximum ionization potential. (Since E_s is fixed, the sync signal amplitude must be small to make the value of m small.) For the particular case of Fig. 11-8, where n is equal to 2, we find that the m or E'/E_s cannot be greater than approximately 1/7.6. This means that if this ratio is attained, curve *A* will be tangent to the positive slope of the sine wave and it will strike the sine wave at $(E_i + E')$.

For values of sync signal amplitude that make m greater than the previous value, the maximum permissible striking voltage, as pointed out previously, will be less than $(E_i + E')$. In Figs. 11-9A and B are illustrated two graphical constructions showing how this is possible. In both drawings, each sawtooth wave is already synchronized and $n = 2$. In Fig. 11-9A, a

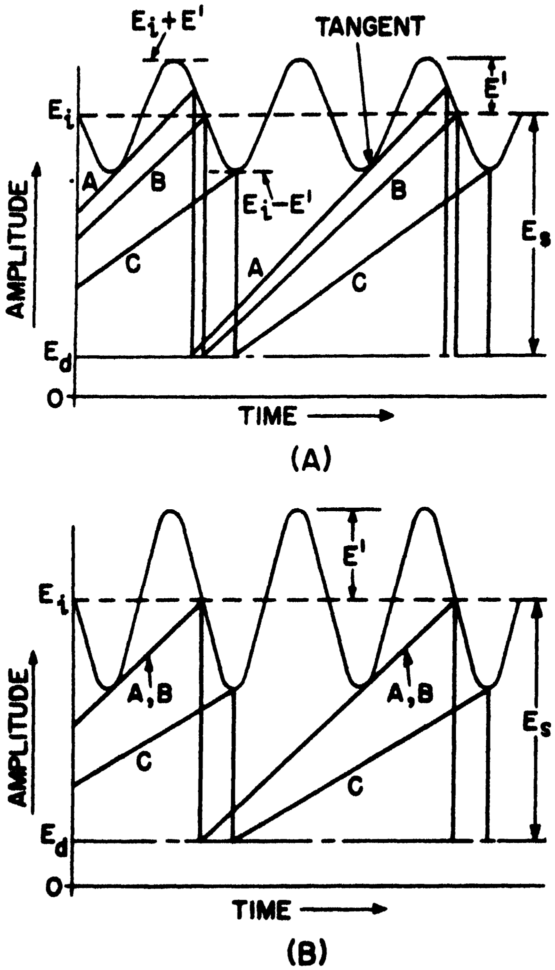

Fig. 11-9.—The strength of the sync signal input helps determine the upper natural-frequency limit of stable synchronization. The limit of graph (A) is higher than that of (B).

strong sync signal has caused the ionization potential to change so that the maximum permissible ionization potential where curve *A* strikes is no longer equal to $(E_i + E')$, but lies in between the potential of $(E_i + E')$ and E_i. Curve *B* in this drawing illustrates the intermediary case where the striking occurs at the ionization potential of E_i and curve *C* represents the minimum possible ionization potential. From a quick examination of curves *A*, *B*, and *C* of Fig. 11-9A, we come to the conclusion that the difference between $f_{n\,(\text{max})}$ and f_o (represented by curves *A* and *B*, respectively) is smaller than $(f_o - f_{n\,(\text{min})})$ (represented by curves *B* and *C*, respectively).

Fig. 11-9B shows a particular case where the amplitude of the sine-wave sync signal is somewhat greater, so that the maximum permissible ionization potential is equal to E_i. In this case, curves *A* and *B* are said to coincide, and the maximum possible natural relaxation-oscillator frequency for synchronization is equal to the required sawtooth frequency; in other words $f_{n\,(\text{max})} = f_o$.

For still stronger values of sync signal, that is, with E' increasing, the maximum permissible striking voltage will be less than E_i, and approaches $(E_i - E')$ as E' tends to become equal to the ionization potential E_i. In cases where the maximum permissible ionization potential lies in between E_i and $(E_i - E')$, synchronization where f_s is twice f_o will occur *only* if the natural frequency of relaxation oscillations f_n is slightly *less than* f_o, or half the frequency of f_s, the sync signal.

You will notice that both curves *A* in Fig. 11-9 are tangent at positive slopes of the sine wave. If a measurement of the ratio of E'/E_s in both drawings is made, it will be found to be greater than $1/(4.3n - 1)$, where $n = 2$, indicating that the maximum permissible ionization potential cannot occur at $(E_i + E')$.

Maximum Free-Running Frequency

From the previous section, we have seen how the strength of the sync signal is the chief factor in determining the maximum permissible striking potential and hence the maximum free-running relaxation frequency under a specific set of conditions. We can graphically illustrate this frequency for any given set of conditions. Let us examine Fig. 11-10. In this drawing, we have plotted an alternating ionization potential superimposed over the fixed potential of E_i. According to the drawing, the sinusoidal curve is represented as $(E_i + E' \sin \omega t)$ and it is plotted against specific values of time indicated by degrees. As pointed out previously, stable synchronization can occur anywhere on the ionization potential curve between 90° and 270°.

In Fig. 11-10, we have drawn a sawtooth wave in such a way that it strikes the negative slope of the ionization curve at point *A* and will also be tangent to the

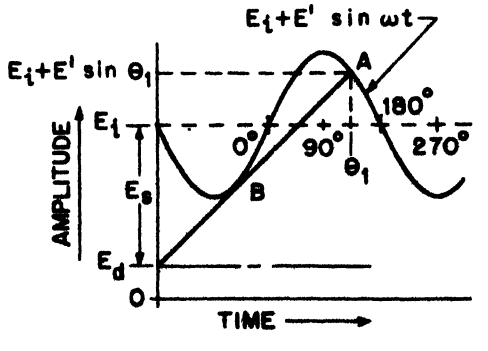

Fig. 11-10.—Graphical illustration of the determination of the upper limit of the natural frequency of relaxation oscillations.

positive slope of the previous half-cycle at point B. We can, therefore, state that the slope of this line AB defines a frequency. We know that this frequency for synchronization must be equal to f_s/n and also that it determines the maximum value of striking voltage at whatever the E'/E_s ratio. For the particular case relative to line AB in Fig. 11-10, let $\omega t = \theta_1$; the maximum striking voltage then is $(E_t + E' \sin \theta_1)$ and the maximum sawtooth amplitude is represented by $(E_s + E' \sin \theta_1)$. From equation (11-7), the maximum natural relaxation-oscillator frequency becomes

$$f_{n\,(\text{max})} = \frac{E_s + E' \sin \theta_1}{E_s} \times \frac{f_s}{n}. \qquad (11\text{-}10)$$

If we omit the subscript from θ_1 we will have a general formula for $f_{n\,(\text{max})}$. Thus

$$f_{n\,(\text{max})} = \frac{(E_s + E' \sin \theta)_{(\text{max})}}{E_s} \times \frac{f_s}{n}. \qquad (11\text{-}11)$$

Equation (11-11) can be rearranged to the more useful form,

$$f_{n\,(\text{max})} = \left(1 + \frac{E'}{E_s} \sin \theta\right)_{(\text{max})} \times \frac{f_s}{n}$$

$$= (1 + m \sin \theta)_{(\text{max})} \times \frac{f_s}{n}. \qquad (11\text{-}12)$$

If we take this latter equation and, for specific values of m, plot n versus θ, we will get a series of curves, one for each different value of m or E'/E_s ratio, which can be used as a ready source for determining $f_{n\,(\text{max})}$. This has been done for six different values of m as seen in Fig. 11-11. As the sync amplitude E' is increased, the ratio of E'/E_s or m is likewise increased. Consequently, from these six curves, we can readily see that as the sync signal is increased, that is, as we advance from curve A to F, the value of n decreases for any particular value of θ.

To determine the value of $f_{n\,(\text{max})}$ at any particular θ setting from the graph, we have to use equation (11-12). The resulting answer will be in terms of f_s, the frequency of the sync signal. For instance, for a θ value of 135°, we intersect curve B at an n value of about 3. The value of m for curve B is 0.125; thus from equation (11-12)

$$f_{n\,(\text{max})} = (1 + 0.125 \sin 135°)\,\frac{f_s}{3}$$

$$= 1.088\,\frac{f_s}{3}$$

$$= 0.363\,f_s.$$

The n value of 3 tells us that 3 cycles of the input signal will appear on the screen. From the above example, it is seen that the maximum natural relaxation frequency is equal to 0.363 times the sync frequency. If f_s is equal to 1,200 cps, then $f_{n\,(\text{max})}$ will equal 436 cps, which is 36 cps above f_o, the synchronized oscillator frequency. If the value of $(1 + m \sin \theta)$ is greater than unity, then $f_{n\,(\text{max})}$ will be greater than f_o; if this quantity is less than unity, $f_{n\,(\text{max})}$ will be smaller than f_o. The graph of Fig. 11-11 can be utilized in other ways, and the example chosen above was one method of illustrating its usefulness.

Zones of Synchronized Operation

It should be remembered that the minimum permissible value of the natural relaxation-oscillator frequency can be easily calculated by the use of equation (11-8); but because of the variation of the maximum frequency with the angle θ, the maximum value of f_n has to be interpreted from the graph of Fig. 11-11 and equation (11-12). The maximum and minimum permissible values of the free-running relaxation frequency for a particular set of conditions have certain *zones of operation* where synchronization is possible. Outside of these zones, synchronization will not occur.

In any problem involving equation (11-12), to obtain the numerical value of $f_{n\,(\text{max})}$ we have to know f_s,

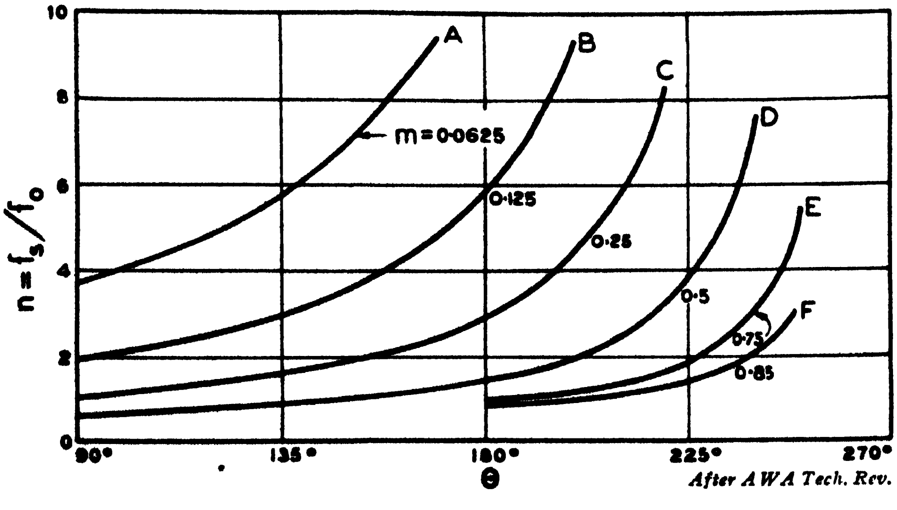

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Fig. 11-11.—For any set condition of m and any value of θ we can determine the value of the maximum natural frequency of relaxation oscillations in conjunction with equations (11-12).

the sync signal frequency. The maximum and minimum values of the ratio f_n/f_s have been plotted against values of m from zero to unity for different values of n from 1 to 5. (Remember that $n = f_s/f_o$.) This is illustrated in Fig. 11-12. The maximum values of f_n/f_s (that is, where $f_n = f_{n\ (\mathrm{max})}$ are indicated by the solid lines and minimum values of f_n/f_s (that is, where $f_n = f_{n\ (\mathrm{min})}$ are indicated by the dashed lines. The regions bounding these maximum and minimum curves are shown shaded for each value of n. *Synchronization, therefore, can occur only within these shaded areas.* From these curves of Fig. 11-12, a number of interesting features are evident:

1. As the value of n is increased, it becomes more difficult to effect synchronization. In other words, the area of the shaded zones decreases as n increases.

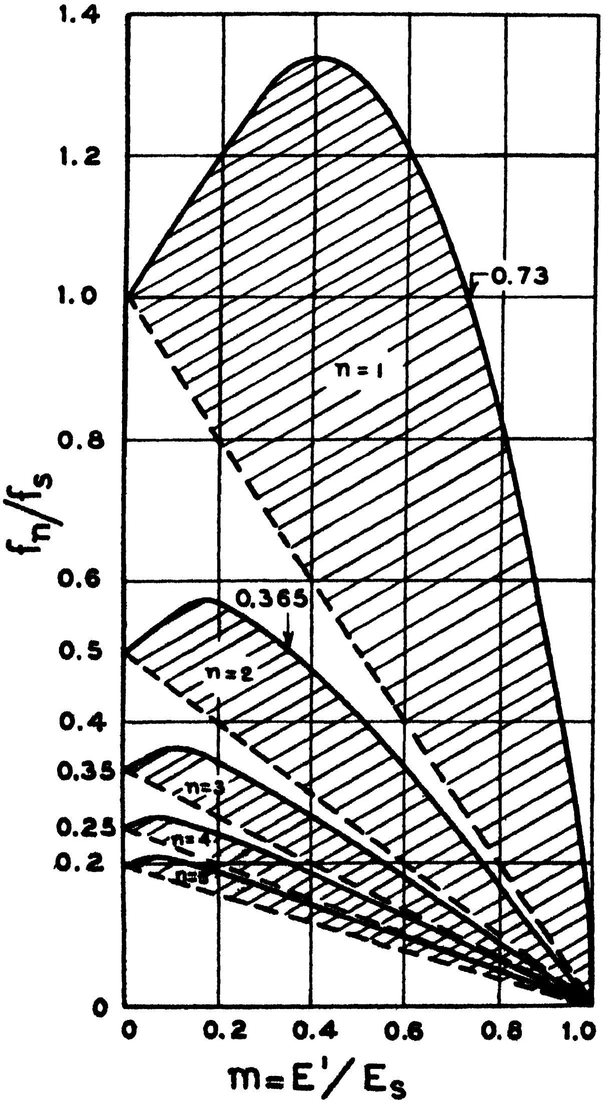

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Fig. 11-12.—Synchronization can occur only in the shaded areas of the drawing.

2. As n increases, the maximum value of natural relaxation oscillations approaches the minimum value.

3. As the sync amplitude increases, and hence as m increases, it becomes more difficult to obtain synchronization. This is even more evident as n increases.

4. When the value of f_n becomes equal to f_s/n, synchronization can only occur for quantities of m that have a maximum value of $0.73/n$. Thus for the case when one cycle appears on the screen, that is, where $n = 1$, m cannot be greater than 0.73. For the case where $n = 2$, f_n/f_s equals 0.5, and we find that m cannot be greater than 0.365. When $n = 5$, f_n/f_s equals 0.2, and the maximum value of E'/E_s is 0.146.

5. For those cases where m is greater than $0.73/n$, the natural relaxation frequency f_n must be less than the synchronized relaxation frequency if locking is to occur.

6. The areas in which synchronization occurs do not overlap for any particular value of m. For every value of m, or for that matter for f_n/f_s, when we go from one synchronized state to another, we pass through regions in which we *cannot* obtain synchronization. This is very important to remember because it also tells us that even though we have a sync signal applied, synchronization is not always possible. For instance, for a fixed ratio of f_n/f_s of 0.4 and when $n = 2$, synchronization will occur only for values of m between 0.2 and 0.5. If the sync signal is too small so that $m < 0.2$, or if it is too large so that $m > 0.5$, synchronization will not be possible under these conditions.

By taking the f_n/f_s or Y axis of Fig. 11-12, multiplying it by n, and then plotting this value of $n(f_n/f_s)$ against different values of m, we have another very convenient means of determining the extent of the maximum and minimum natural relaxation-oscillator frequencies. A graph of this plot is illustrated in Fig. 11-13. Since $n = f_s/f_o$, then $n(f_n/f_s)$ simplifies down to f_n/f_o. Therefore, the graph of Fig. 11-13 is actually a plot of the ratio of the free-running frequency of relaxation oscillations to the synchronized frequency of relaxation oscillations versus m.

The plot is made for values of n between 1 and 10. The solid lines indicate the maximum values of f_n/f_o for any value of m from zero to unity. The single dashed line indicates the minimum value of f_n/f_o and is seen to be the *same for all values of n*. This feature of the graph makes it very useful for interpretation of various factors concerning synchronization. Below we have listed a number of interesting features which are evident from this graph:

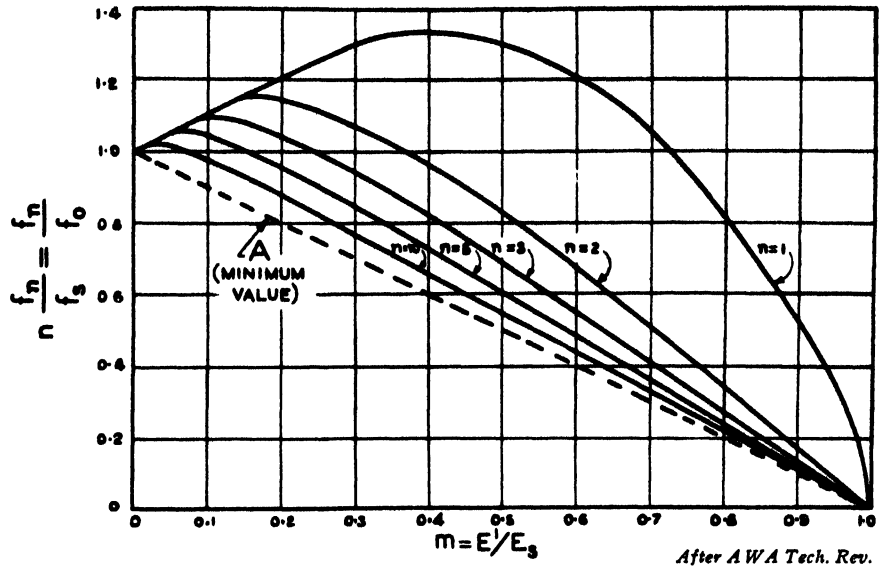

Fig. 11-13.—A plot of the ratio of natural frequency to the synchronized sawtooth frequency versus different values of *m*.

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1. As *n* increases, the area bounded by the maximum value of the *n* curve and curve *A* decreases. This indicates that the extent of synchronization decreases as *n* increases.
2. As *n* increases, the value of *m* (and hence the sync signal amplitude) must be very small in order to have $f_n > f_o$. In other words, the area of synchronization where $f_n/f_o > 1$ decreases as the value of *n* increases.
3. In general, the area of synchronization above the line where f_n/f_o is equal to 1 indicates the possible plane of synchronization where $f_n > f_o$, or where $f_n > f_s/n$. This area of synchronization decreases as *n* increases.
4. As the number of cycles to appear upon the screen is increased (i.e., an *n* increases), it is more difficult to effect synchronization. This is so especially when $f_n > f_o$.
5. The complete area bounded by the $n = 1$ curve and curve *A* is where synchronization can *only* occur. For all values of f_n/f_o and *m* that intersect outside this region, synchronization is not possible.
6. As *n* increases to infinity, the maximum permissible natural relaxation frequency approaches the minimum permissible relaxation-oscillator frequency. This is indicated by the curve for $n = 10$ approaching curve *A*.

The analysis thus far was based upon the fact that the sawtooth curves are ideal, that is, they have linear traces and zero retrace times. Although this is only ideal, in actual practice, the curves of Figs. 11-12 and 11-13 are not very far off the practical cases and are well within the usual allowable limits between theory and practice. Although sine waves were used as the sync signal, any other periodic function would have the same effects relative to synchronization. That this is so is witnessed by the fact that whatever type of signal, which is observed on the screen of the cathode-ray tube, is also used as the sync signal of the oscilloscope.

Apparent Synchronization

We have previously pointed out that the synchronized relaxation-oscillator frequency has to be equal to, or some integral multiple of, the sync signal, for synchronization to be possible. This frequency relationship is indicated by the expression $f_o = f_s/n$, where *n* can be any integer including unity. Every successive synchronized relaxation oscillation, that is, *every cycle of the synchronized sawtooth wave, was identical in frequency and amplitude.*

We can establish this *identity* as the *rule* for synchronization. This can be checked easily by examination of the previous synchronization patterns (see Figs. 11-5 and 11-6). There are, however, cases where this identity does exist but does not truly represent synchronization or locking of the sawtooth oscillator with the sync signal. These cases are, therefore, exceptions to the rule. For *all* these exceptions, the sawtooth signal during application of the sync signal *must* strike the ionization potential curve at E_i for every successive relaxation oscillation. That is, when that part of the ionization potential, represented by $E' \sin \omega t$, is equal to zero, striking occurs, and this happens when ωt is equal to 0° or 180°. Under this condition, the natural frequency of sawtooth oscillation must be equal to the frequency of the sawtooth signal during the application of the sync signal.

You will note that we did not refer to the sawtooth signal, when under the influence of the sync signal in this particular case, as a synchronized sawtooth signal (i.e., as f_o). This is so because, as we mentioned, synchronization does not actually exist when the firing constantly occurs at the potential E_i. To indicate that there is a difference between this synchronized and unsynchronized sawtooth, we will call the relaxation oscillations when under the influence of the sync signal (in this exception to the rule), by the symbol f_o'. Therefore, for the case under discussion

$$f_n = f_o'. \qquad (11\text{-}13)$$

Since the striking voltage, when equal to E_i, occurs twice during each cycle of sync signal, a restriction is, therefore, placed upon f_o'. We can represent this restriction as

$$f_o' = \frac{2f_s}{p} \qquad (11\text{-}14)$$

where p can have any integral value including unity.

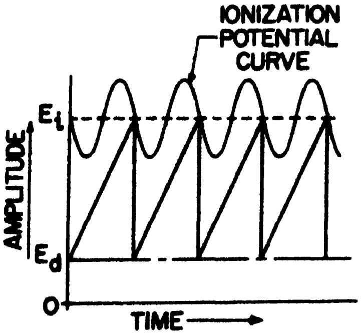

Fig. 11-14.—Representation of apparent synchronization where the natural sawtooth frequency is equal to the sync signal frequency and fires at the ionizing potential of the tube that would exist without any sync signal applied.

Now let us examine Fig. 11-14. This drawing shows us the case where the frequency of the sawtooth signal f_o' is equal to that of the sync signal f_s. In other words, for Fig. 11-14, p of equation (11-14) is equal to two. If the value of p is always higher than two, f_o' will always be less than f_s. Note that each cycle of the sawtooth signal strikes the sine curve at potential E_i.

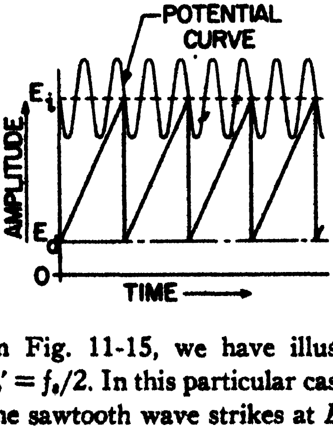

Fig. 11-15.—A similar case of apparent synchronization but where the sync signal frequency is equal to twice the natural sawtooth frequency.

In Fig. 11-15, we have illustrated the case where $f_o' = f_s/2$. In this particular case, p is equal to four and the sawtooth wave strikes at E_i every second cycle of the sine curve. If you will glance at Figs. 11-14 and 11-15 once again, you will note that the sawtooth strikes the sine wave along negative slopes of the curve. If the sine waves were each shifted 180°, the sawtooth signals would strike at E_i but along the positive slopes of the sine curve. Although we generally consider the positive slope a point of unstable synchronization, in this particular case, it cannot be considered as such, since we actually do not have any locking effect when the striking potential is constantly at E_i.

In the previous two drawings, the value of p was an even integer, being 2 for Fig. 11-14 and 4 for Fig. 11-15. In the first case, $f_o' = f_s$, and in the latter, $f_o' = f_s/2$. Such ratios can also exist in the true case of synchronization (that is, where the ionization potential is hit at the same point other than at E_i during each successive cycle of the sawtooth), where we have $f_o = f_s$ and $f_o = f_s/2$. We commonly refer to *all* cases where the frequency of the relaxation oscillator is less than that of the sync signal by an integral value as *frequency division*. By this we mean that we have to *divide* into the frequency of the sync signal by some integer greater than one to obtain the frequency of relaxation oscillations. In these cases, the division is by a whole number. For *simple frequency division* (that is, where f_s is an integral multiple of f_o') to hold for equation (11-14), p has to be even and greater than two.

Let us examine the situation of equation (11-14) when p is not an even number but is an odd number greater than one. If p is equal to 3, equation (11-14) becomes

$$f_o' = \tfrac{2}{3} f_s$$

which means that the frequency of relaxation oscillation must be ⅔ that of the sync signal. This particular situation is illustrated in Fig. 11-16, where curve *A* represents the sawtooth signal having ⅔ the frequency of the sine curve. This is often called *fractional division*. By rearranging equation (11-12) as follows,

$$f_o' = \frac{f_s}{\tfrac{3}{2}}$$

we can actually see that f_s is divided by a fractional number. An interesting feature of fractional division is that the sawtooth strikes the sine wave alternately on the negative and positive slopes of the curve at E_i.

Fractional division cannot occur when actual synchronization exists. For fractional division to occur in this case, the number n in the relation $f_o = f_s/n$ would have to be a fraction, which we know to be impossible.

From the drawings of Figs. 11-14, 11-15, and 11-16, we see that the sync signal has no effect upon the re-

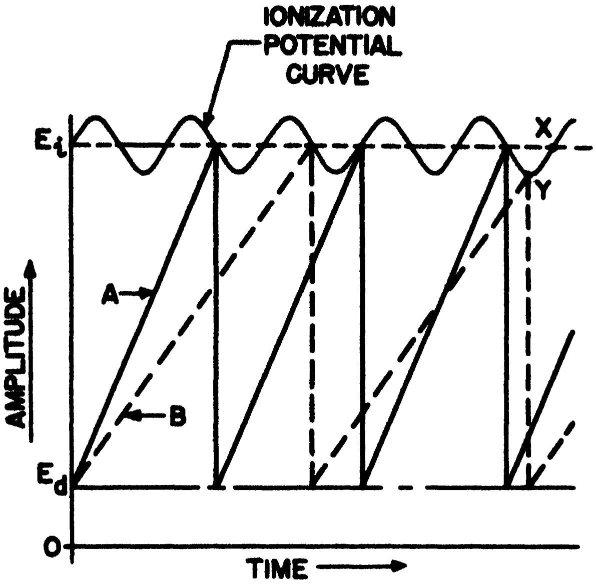

Fig. 11-16.—Illustration of simple frequency division. For curve *A* the sawtooth frequency is ⅔ the sync signal frequency; for curve *B* the frequency ratio is initially ⅖ but its amplitude is such that it strikes the ionization curve prematurely.

laxation oscillations because each cycle strikes at ionization potential E_i, which would also exist if the sync signal were not present. However, if the amplitude of the sync signal is too strong, the sawtooth wave may intercept the ionization potential curve at a premature value less than E_i. Curve *B* of Fig. 11-16 illustrates such a case. This curve initially started out as a case of fractional division where p is equal to five, that is, where $f_o' = 2f_s/5$. Thus each cycle of the sawtooth is supposed to strike the sine curve at E_i every fifth half-cycle. However, when the second cycle of sawtooth starts to rise, it heads for point *X* on the E_i line, but due to the strong sync signal, it strikes the sine curve at point *Y*. When this happens, the capacitor discharges and we have a complex situation where each successive cycle is different in frequency and amplitude.

From the analysis thus far, we can readily visualize what will happen if p is greater than 5. Of course, if we want to avoid the effect of hitting the sine curve other than at E_i, the strength of the sync signal has to be small. This is even more so as p becomes greater, that is, as f_o' becomes much smaller than f_s.

What about situations where the frequency of the sawtooth signal, whether it be represented by f_o or f_o', is greater than f_s? Such cases cannot be synchronized, which means that the situation where f_o is a multiple of f_s is not possible. However, it is possible to have f_o' greater than f_s, but only for *one* frequency. Examination of equation (11-14) will reveal that this single case is where p is equal to unity. In other words, only when $f_o' = 2f_s$ is this situation possible.

In Fig. 11-17 a graphical analysis appears, showing this case. Note that each cycle of sawtooth strikes each half-cycle of the sine wave at E_i. This situation is often referred to as *frequency multiplication;* it can exist only where the sawtooth constantly strikes at E_i and, therefore, does not represent a true case of synchronization.

The existence of conditions where the striking voltage must be equal to E_i for every cycle of sawtooth establishes the fact that the natural frequencies of relaxation oscillations f_n must always be equal to f_o'. Since f_o' is characterized by the relation $2f_s/p$, where p can be any integer, there is no limiting condition. If we were to gradually reduce the sync signal and then remove it, striking would still occur at E_i. Thus, this does not represent the case where the sync signal causes the sawtooth to become locked-in, although the pattern may appear stationary on the screen. That we cannot consider this as a case of synchronization is evidenced by the fact that there is no finite value of E' to substitute in the relation E'/E_s (the ratio m), and so these particular cases do not fall into any of the regular zones of synchronization of Fig. 11-12 or within the limits of the curves of Fig. 11-13.

Oversynchronization

It is very important to know how to operate your oscilloscope properly. This is especially so in the use of the sync control because the strength of the sync signal affects the relaxation oscillations. Although you may not realize it by the picture you are viewing on the screen of your oscilloscope, the reproduced picture may be distorted if too much sync signal is applied. This

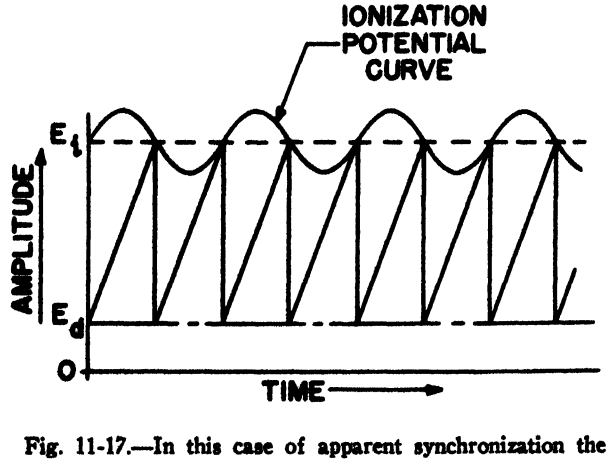

Fig. 11-17.—In this case of apparent synchronization the sawtooth frequency is equal to *twice* the sine-wave frequency.

was indicated in Chapter 10. We mentioned previously that for proper oscilloscope operation, the fine sweep-frequency control has to be adjusted first, keeping the sync control at zero, to stop, as best as possible, the pattern on the screen. *After* this adjustment is made, some sync signal is injected into the sweep circuit by careful manipulation of the sync control. As soon as the pattern is "stopped," no more sync signal should be injected.

If the amplitude of the sync signal is too great, it may cause the sawtooth sweep to intercept the ionization potential curve at *different* values and thereby each successive cycle of the resulting sawtooth will not have the same frequency and amplitude as the preceding one. Thus the sweep signal itself would be distorted. Naturally, a sweep signal such as this would not reproduce the input signal in its true form.

In Figs. 11-12 and 11-13, we illustrated zones of synchronized operation for various different values of *n*. We investigated most of these cases that fall within these zones, that is, within the shaded areas of Fig. 11-12 and within the area enclosed by the curve $n = 1$ and the dashed curve of Fig. 11-13, but we have not said anything about what occurs outside these areas. The application of too much sync signal may cause the operation of the circuit to fall outside of these areas. When this happens, we say that we have a condition of *oversynchronization.*

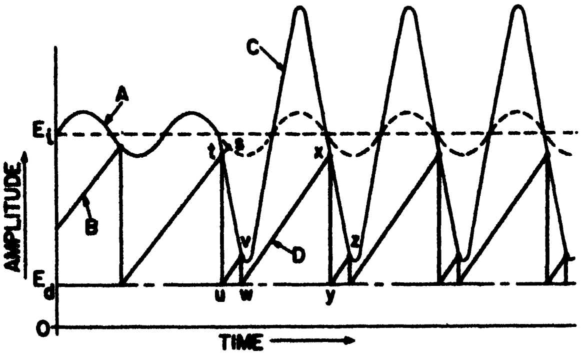

Fig. 11-18.—Graphical analysis of oversynchronization. Note the shape of the sawtooth wave during the time the sync signal has increased.

Let us graphically analyze what happens to the relaxation oscillations because of oversynchronization and then illustrate some practical cases by oscillograms. In Fig. 11-18, sine curve *A* represents a normal sync signal and sawtooth wave *B* the accompanying relaxation oscillation. The relationship between the frequency of the sync and sweep signal is $f_s = f_o$, so that only one cycle of sine wave will appear on the screen of the cathode-ray tube. Assume that shortly afterwards the strength of the sync signal is raised to the point where the resulting ionization potential takes the form illustrated by sine wave *C*.

As the sawtooth wave rises, it would have originally headed for point *s* on the original curve of ionization potential. However, due to the increase in the alternating ionization potential, it strikes the new curve at point *t*. The capacitor discharges at this latter point, and when the tube reaches the deionization potential at point *u*, the capacitor starts to charge up again at the same rate as before the increase in sync signal. The sawtooth signal rises again until it reaches the ionization curve, sooner than expected, at point *v*. The capacitor then discharges to point *w* and charges up again at the same rate until it strikes the ionization curve at point *x*. Points *t* and *x* are at the same potential, as are points *v* and *z*. For the same amount of increased sync signal, the relaxation oscillations assume the shape illustrated in the drawing after point *u*.

You will note that with the oversynchronization illustrated in Fig. 11-18, the frequency of the relaxation oscillations has been doubled and is equal to twice that of the sync signal. That is, two cycles of sawtooth signal appear for every sine-wave cycle. Under this condition one cycle of sawtooth will no longer sweep the screen of the cathode-ray tube; the nonuniform sawtooth will cause the pattern to be distorted.

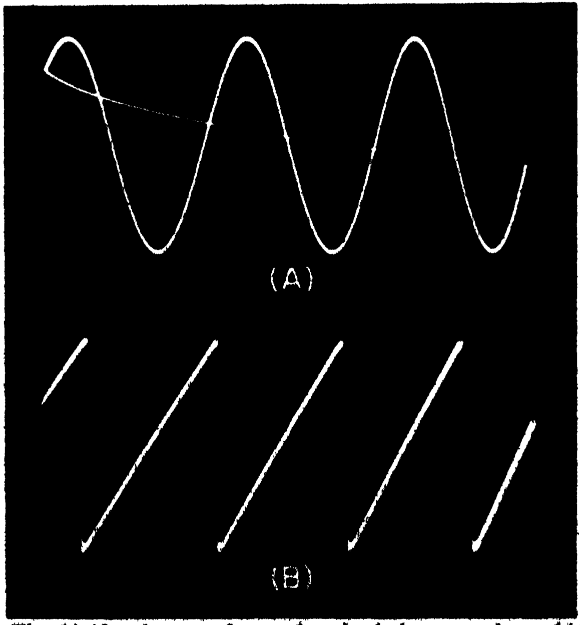

Fig. 11-19.—A properly synchronized sine wave is at (A) and (B) is the synchronized sawtooth causing this sine wave.

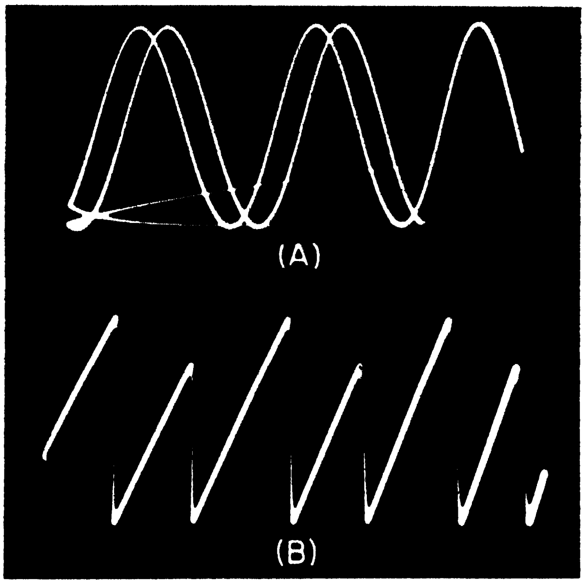

Fig. 11-20. — A small amount of oversynchronization changed the sine wave of Fig. 11-19A to (A) of this figure. The sawtooth at (B) is that causing the pattern at (A).

Now let us examine some typical oscillographic waveforms of operation under different amounts of sync signal. In the following drawings, the input signal is a 5,000-cps sine wave which measures about 100 volts peak to peak, unless otherwise indicated. In Fig. 11-19 appear two photos. That at (A) illustrates three cycles of a sine wave, *properly* synchronized, and that at (B) the synchronized sawtooth signal causing the picture at (A). Since three cycles of sine wave appear on the screen, $n = 3$; hence $f_s = 3f_0$. Using these two photos for comparison, we will gradually keep increasing the strength of the sync signal and observe the resulting pattern of the input signal and sawtooth sweep.

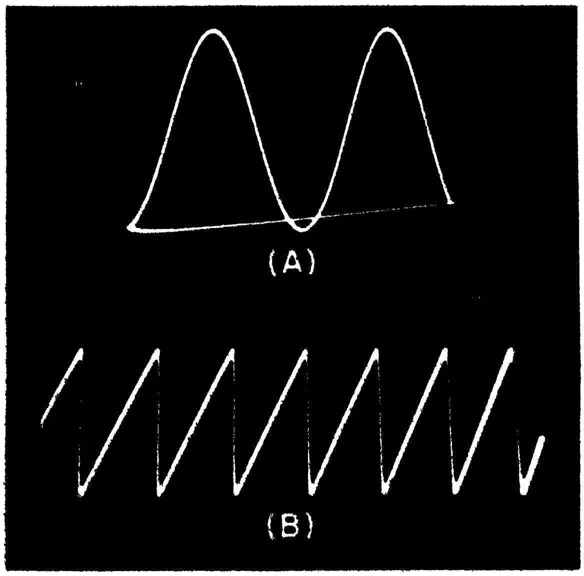

Fig. 11-21.—The sync signal for this figure was increased a little more than that for the latter one. The sine wave at (A), although not appearing distorted, is decreased to two cycles and the sawtooth at (B) is reduced in amplitude and increased in frequency.

The two patterns of Fig. 11-20 are the result of a small amount of oversynchronization. That at (A) indicates the resulting pattern, while that at (B) is the relaxation oscillation causing the pattern at (A). Note that the sawtooth wave at (B) closely resembles the theoretical drawing of Fig. 11-18, except that in Fig. 11-20B the second sawtooth cycle does not have as small an amplitude as that of Fig. 11-18.

For Fig. 11-21, the sync signal was increased still further. The photos represent an interesting phenomenon in that the sine wave at (A) and the sawtooth signal at (B) are not distorted. The increased sync signal at this point has changed the shape of the ionization-potential curve in such a way that each cycle of the resulting sawtooth signal is of the same frequency and amplitude. The repetition frequency of this sawtooth wave is such that only two cycles of the input signal appear on the screen instead of three cycles as in Fig. 11-19. This means that the frequency of relaxation oscillation for the signal at Fig. 11-21B is only one-half that of the input signal ($f_o = f_s/2$). In this particular case, too much sync signal has not caused distortion, but has only increased the frequency and decreased the amplitude of the sawtooth signal.

If the sync signal is increased still further, we find that we once again distort the pattern. This is shown by the oscillograms of Fig. 11-22. The distorted sine

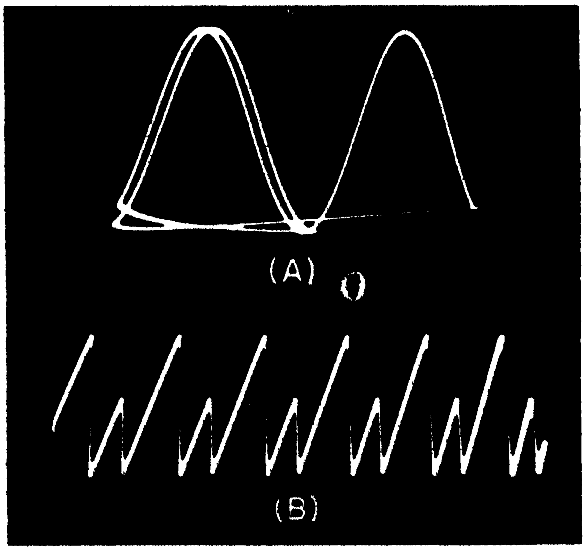

Fig. 11-22.—A further increase in sync signal causes the distorted sine wave at (A) due to the distorted sawtooth at (B).

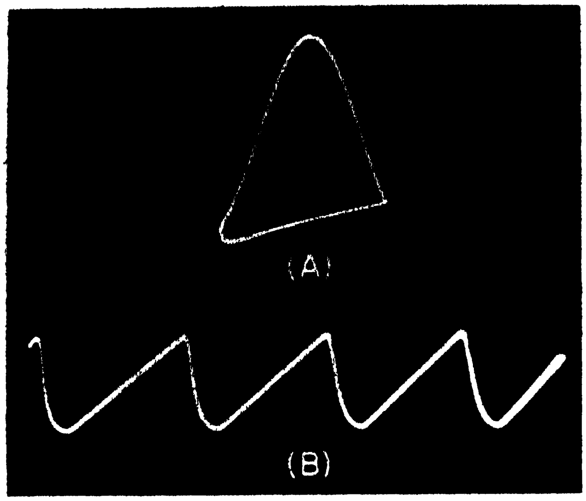

Fig 11-23.—Increasing the sync control to its maximum position results in the distorted sine wave at (A) due to the distorted sawtooth at (B).

wave appears in part (A) and the distorted sawtooth in part (B). Comparing this sawtooth of Fig. 11-20B with that of Fig. 11-22B, we see that they are somewhat similar in shape, but that in the latter, the frequency is increased and the amplitude is decreased.

When the sync control of the oscilloscope in use is advanced to its maximum point, the photos of Fig. 11-23 result. That at (A) is the reproduced pattern of the input signal and that at (B) the distorted sawtooth that caused the pattern at (A). In order to find out what a stronger amount of sync signal would do to the relaxation oscillations, the 5,000-cps input signal to the oscilloscope increased from 100 to 300 volts peak to peak with all controls remaining unchanged from what they were to produce Fig. 11-23. The pattern of the reproduced input signal appears in part (A) and the distorted sawtooth causing this pattern is shown in part (B) of Fig. 11-24. Compare these pictures with those of Fig. 11-23. In considering the photos of Fig. 11-24, remember that too high an increase in the strength of the input signal may cause overloading and distort the sine wave before it reaches the cathode-ray tube.

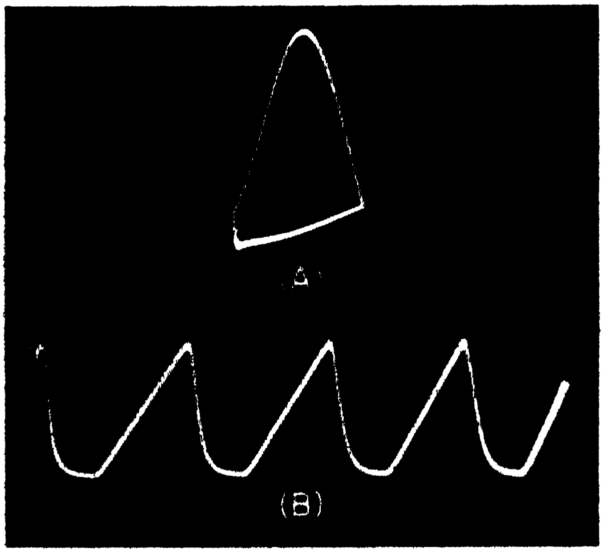

Fig. 11-24.—The strength of the sine wave was increased over that for the latter figure and keeping all controls as they were previously, the resultant wave is shown at (A) and the sawtooth at (B).

When the input sine wave is changed to 250 cps at a peak-to-peak voltage of 66 volts, the resulting patterns due to oversynchronization are illustrated in Fig. 11-25. Note the odd shape of these curves. That at (A) represents the reproduction of the input signal, while that at (B) is the distorted sawtooth that reproduced the pattern at (A). The pattern of the sweep signal is such that we see two small relaxation oscillations in between the large oscillations. It is these small sawtooth waves that cause the "zig-zag" distortion on the left side of the pattern in Fig. 11-25A.

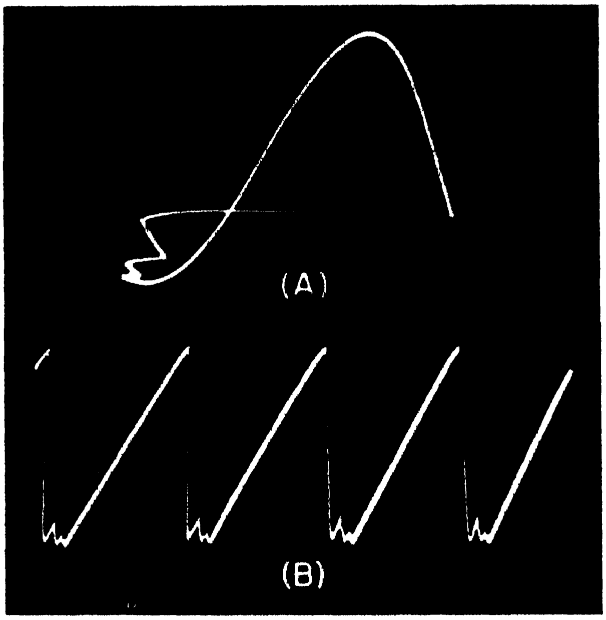

Fig. 11-25.—The frequency of the input sine wave was changed from 5,000 cps to 250 cps and at a peak-to-peak voltage of 66 volts. The resultant sine wave is at (A) and the sawtooth causing it at (B).

A graphical analysis of how the sawtooth signal of Fig. 11-25B can take on the shape shown is illustrated in Fig. 11-26. With too much sync signal applied, the ionization potential appears as shown in the drawing. As the sawtooth wave rises, it strikes the ionization curve at point *a* and the capacitor then discharges. As the sawtooth starts to rise again at the same rate as

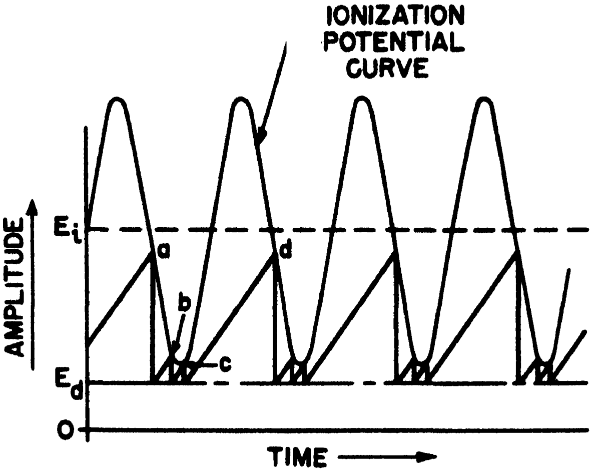

Fig. 11-26.—Graphical analysis of how oversynchronization can cause the distorted sawtooth wave of Fig. 11-25B.

before, it does not have a chance to rise very much and it strikes the sine wave at point *b;* the capacitor then discharges again. The same thing happens all over again with the sawtooth on the next rise time hitting the sine curve at point *c* with the capacitor then discharging. What has happened is that for every cycle of sync signal, we have three cycles of sawtooth, each individual sawtooth cycle differing in frequency and amplitude. The interesting thing about this drawing is that the sawtooth signal between points *a* and *d* is successively repeated.

Increasing the sync signal to its maximum when its input is still 250 cps at 66 volts results in the oscillograms of Fig. 11-27. That at (A) is the reproduced signal of the input and that at (B) the distorted sawtooth causing the picture at (A). Looking at Fig. 11-26 once again, you can easily visualize that a further increase in the sync signal, and hence ionization potential, will cause the negative peaks of sine wave to go below the deionization potential. This is what happened in the case for Fig. 11-27, and the very strong sync signal has actually blanked out some of the sawtooth signal.

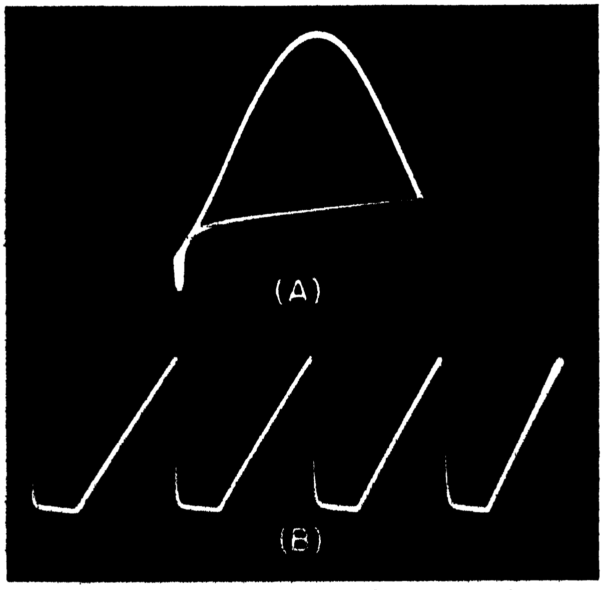

Fig. 11-27.—A maximum increase in the sync signal over that of Fig. 11-25 results in the reproduced picture at (A); the distorted sawtooth is at (B).

SYNCHRONIZATION OF MULTIVIBRATORS AND BLOCKING OSCILLATORS

The analysis of synchronization in multivibrators is, in general, more complex than in the case of the simple thyratron relaxation oscillator. However, the qualitative analysis relative to the thyratron synchronization applies to multivibrators also.

Part of the input signal is also used as the control or synchronizing voltage for multivibrator circuits when employed as sweep circuits in oscilloscopes. This control voltage is applied to one of the grids of the multivibrator. In the following discussion, for the sake of illustration, this control or sync voltage will be assumed sinusoidal unless otherwise indicated. Since the sync signal is alternating, the instantaneous bias on the grid of the tube to which it is applied will also be changing. The effect of this is that the sync signal will actually control the start of the conduction period of the tube. From what has already been discussed about multivibrators in Chapter 9, we can readily understand

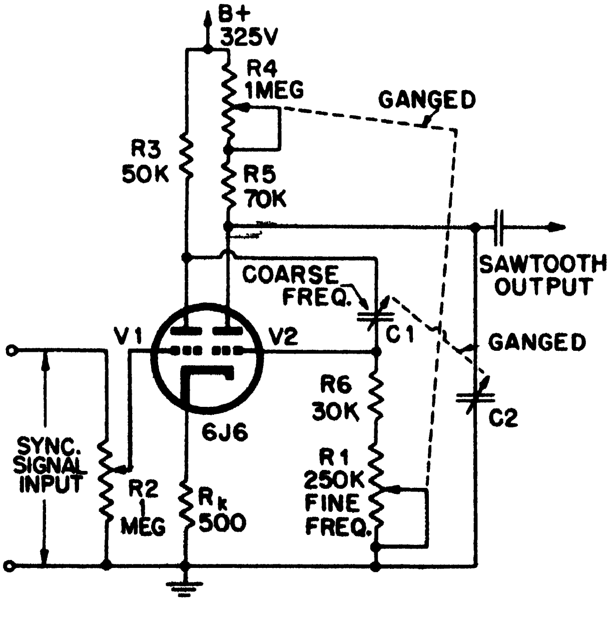

Fig. 11-28.—A typical cathode-coupled multivibrator circuit showing the application of sync voltage.

that controlling the conduction period of one tube will automatically control the operation of the circuit.

With multivibrators, it is possible to have the frequency of the applied sync signal f_s equal to an integral multiple of f_o, the synchronized frequency of relaxation oscillations. In other words, the ratio of f_s/f_o for multivibrators can be equal to, or greater than, unity. Under special applications, as will be seen later, the f_s/f_o ratio may be less than unity. A typical multivibrator circuit as used in oscilloscopes, showing the circuit of sync signal input, is illustrated in Fig. 11-28. (This circuit is the same as that of Fig. 9-67, except for the sync input.) The sync signal input is applied across potentiometer *R2*, which is used to vary the amount of sync input to the grid of *V1*.

Sine-Wave Synchronization of Multivibrators

Let us now graphically analyze how a sine-wave control signal affects synchronization of a multivibrator circuit similar to Fig. 11-28. In this analysis, we will use the grid-voltage waveform of the multivibrator to illustrate synchronization. This is indicated in Fig. 11-29. That part of the grid-voltage curve between times *A* and *C* represents that waveshape without any sync signal applied. At time *C*, a sinusoidal sync signal is applied, and instead of the capacitor discharge curve being a simple exponential curve as in the previous cycle, it actually becomes a sinusoidal curve superimposed over the exponential curve. The dashed curve represents the grid voltage if no sync signal were applied, while the solid line is the actual curve.

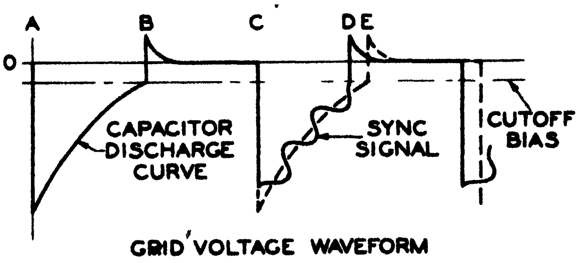

Fig. 11-29.—Analysis of how a sine-wave signal is used as the sync voltage when applied to the grid of a multivibrator.

The sync signal controls the time when the voltage on the grid of the tube in question will reach the cutoff bias and, therefore, cause the tube to conduct. Note how the sync signal of Fig. 11-29 has caused the free-running frequency of the multivibrator to increase by becoming locked-in with the sync signal. In the example illustrated, the frequency of sync signal is three times as great as the frequency of the synchronized *half-cycle* of multivibrator action, thus $f_s = 6f_o$. This frequency relationship is, of course, very simple to see when the multivibrator is symmetrical. This also means that the instantaneous voltage on the grid of each tube is exactly the same at the start of each individual half-cycle.

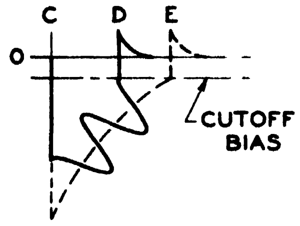

Fig. 11-30.—An increase in the sync signal over that for the previous figure causes the tube to reach its cutoff point earlier and hence increases the frequency of the synchronized multivibrator.

If the strength of the sync signal is increased, the cutoff point of the tube will be reached at an earlier instant. In other words, with an increase in sync amplitude, the synchronized frequency will be further increased, but the ratio of f_s/f_o will be decreased. This can be seen better from Fig. 11-30. This picture depicts an increase in the strength of the sync signal of Fig. 11-29 so that after the *second* cycle of sync signal, the cutoff bias of the tube is reached and conduction starts. The dashed curve is the grid-voltage waveform that would exist if no sync signal were applied; it is the same curve as that between times *C* and *E* of Fig. 11-29. With this fact understood, you can readily see that the period of the synchronized half-cycle of multivibrator grid voltage between times *C* and *D* of Fig. 11-30 is less than that between times *C* and *D* of Fig. 11-29. Also note from Fig. 11-30 that the effective sync frequency has been reduced. Consequently, it is evident that an increase in sync amplitude will greatly reduce the f_s/f_o ratio; the ratio in Fig. 11-30 is 4 to 1, as compared to a ratio of 6 to 1 in Fig. 11-29.

If the multivibrator circuit uses a positive grid-bias return (as in some of the circuits analyzed in Chapter 9), the stability of synchronization is increased because the discharge curve over which the sync signal is superimposed is more linear (see Fig. 9-75B). Hence, a greater increase in sync amplitude is required in order that the f_s/f_o ratio change.

From the analysis thus far, we see that the frequency of the sync signal can be much greater than that of the multivibrator because it is only that small part of the sync signal near the cutoff bias of the tube that affects the oscillation. Besides the amplitude, the frequency of the sync signal also affects the synchronized frequency of multivibrator oscillations. A slight change in the frequency of the sync signal will bring about a small change in the synchronized frequency of multivibrator oscillations. This is indicated in Fig. 11-31. Part (A) of this drawing shows two cycles of sync signal superimposed over the grid-voltage discharge curve. The phase relationship between the sync signal and grid voltage in this drawing is such that the sync signal reduces the grid voltage to the cutoff value of the tube at the same time as the exponential discharge curve reaches the cutoff value. The time for the completion of the two cycles of sync signal and also the half-cycle of grid voltage is represented by time *F* to *G* in part (A) of the drawing.

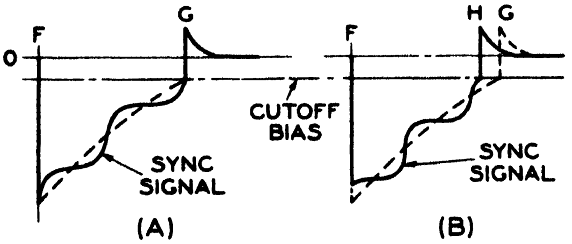

Fig. 11-31.—The sine-wave sync signal of part (A) when slightly increased in frequency causes an increase in the synchronized frequency of multivibrator oscillations as shown at (B).

Assume now that the frequency of the sync signal is increased by a small amount. This is indicated in part (B) of Fig. 11-31 by a reduction in the *time* required to complete the two previous cycles of sync signal. For synchronization to be effective at the end of exactly two cycles of this increased sync signal frequency, there must be a *phase* change between the beginning of the grid-voltage curve and the sync voltage. This phase change is evident in part (B), and it causes the negative amplitude of the grid voltage at the beginning of the curve to decrease from what it was at part (A). Glancing at these two drawings, we see that the time required to complete the half-cycle of grid voltage is decreased from *FG* to *FH,* indicating that the frequency of the grid voltage is increased and is made to lock-in with the sync signal.

You will note that in the drawings of Figs. 11-29, 11-30, and 11-31, there is no sinusoidal sync signal indicated on that part of the grid voltage above the zero voltage line. This is so because as soon as the tube to which the sync voltage is applied starts conducting, the grid is driven positive and grid current starts to flow. It flows only for a short time, however, and the grid is at zero voltage during the conduction period of the tube. Thus any sinusoidal sync voltage would cause grid limiting, which in effect prevents the sync voltage from being additive to the grid voltage at those times. Another important factor in the operation of the multivibrators as indicated in these same drawings is that synchronization will only occur when the sync frequency is higher than the free-running or natural frequency of the multivibrator.

The first cycle of the grid voltage subject to the sync signal usually will not be the final synchronized waveshape as is indicated in Fig. 11-29. The exact shape of this first cycle is, as pointed out by Fig. 11-31, dependent upon the frequency and phase of the sync signal at the time it is applied to the circuit. After application of the sync signal, a few cycles of operation must normally elapse before the multivibrator will fall into a synchronized steady-state condition in which every one of its following cycles will be identical.

The discussion thus far was based upon that of a symmetrical multivibrator, but, as we know, an asymmetrical multivibrator is the type that is used to produce a sawtooth output waveform. However, whatever was said concerning the superposition of the sync signal to the grid-voltage curve as in Figs. 11-29 and 11-30 in controlling the frequency of operation of symmetrical multivibrators applies equally well to unbalanced types. The primary difference is that the relative ratios of f_s to f_0 vary in accordance with the degree of unbalance. The tendency to synchronize at different f_s/f_0 ratios is increased for asymmetrical multivibrators. It becomes too complex to graphically illustrate various f_s/f_0 ratios for different degrees of asymmetry in a multivibrator.

Of course, we have only referred to synchronization of multivibrators as normally employed in oscilloscopes. In these circuits, the sync signal is usually fed to the grid circuit of one tube. There are, however, numerous other methods of introducing the sync signal, and in accordance with these methods of sync-signal introduction and the symmetry of the circuit, the multivibrator can be made to favor special f_s/f_0 ratios or to show no particular distinction at all. For further details about such circuits, consult the bibliography.

Multivibrator Synchronization by Pulses

Thus far, we have indicated synchronization of thyratrons and multivibrators by assuming the sync signal to be a sine wave. As mentioned, other types of sync signals could have been used. If fact, synchronization of multivibrators by pulses of short time duration is generally more satisfactory. The pulses may be either positive or negative, and there are numerous ways in which they can be applied to the circuit. The use of pulses as synchronizing signals is analogous to the use of pulses for triggered or driven types of multivibrators. These pulses are applied directly either to the grid or to the cathode circuit of the tube in question. Whatever the polarity of the signal applied to the cathode, it will effectively appear in the grid circuit of opposite polarity.

Let us first consider the case where positive pulses are effectively superimposed on the grid voltage of a symmetrical multivibrator. As in the previous discussions on synchronization, it is the amplitude and frequency of the sync pulse signal which will determine whether or not the pulse will be able to synchronize the multivibrator properly. In Fig. 11-32, we have illustrated a case of synchronization by the use of a positive pulse superimposed on the grid-voltage waveform. You will note that the positive pulses are effective on *both* half-cycles of the grid-voltage waveform. However, those positive pulses that cause the grid voltage to increase above its zero value, as pulses *2* and *3* in the drawing, will cause only a *momentary* increase in grid and plate current. It will have no effect on the action of the multivibrator when it is thus applied to the conducting tube.

Pulse number *1*, which is effective during the non-conduction period of the tube to which it is applied, is of insufficient amplitude to increase the grid voltage to cutoff. When pulse *4* is effective, the phase relationship between the grid voltage and the pulse is such that the amplitude of the pulse is then sufficient to cause the grid voltage to rise above its cutoff value, and thereby make the tube conduct at time *D*. Time *E* in the drawing indicates where the free-running frequency of the multivibrator would have made the tube conduct. After this initial point (time *D*) has been reached, the sync pulses that follow will each cause the tube to become conductive at a time sooner than if no sync signal were applied. After a small period of time, the pulses will cause conduction of the tube at exactly the same point during each cycle of grid voltage, and thus synchronization is complete.

If you will glance at the drawing of Fig. 11-32 once again, you will see that the synchronized grid-voltage waveforms at the extreme right of the drawing are no longer symmetrical. Therefore, we find that when a sync signal is applied to one grid of a symmetrical multivibrator, it will not only tend to change the frequency of the multivibrator to effect synchronization, but will also make the resulting waveform asymmetrical for each cycle.

In the analysis of Fig. 11-32, the sync-pulse frequency (i.e., the pulse repetition rate) is slightly greater than the free-running frequency of the multivibrator. The sync signal causes the multivibrator frequency to become locked-in with pulse frequency. Thus, f_s is equal to f_0. If the amplitude of the pulse is not strong enough, it may be difficult to effect synchronization. This is best illustrated by the diagram of Fig. 11-33. In this drawing, the frequency of the sync signal is still slightly greater than the free-running frequency of the multivibrator, but the sync pulses are of insufficient amplitude to effect synchronization. In other words, it is difficult for the sync pulse to cause the grid voltage of the tube to which it is applied to reach the cutoff value sooner than it normally does.

It is also possible to synchronize the multivibrator by sync pulses that are a multiple of the synchronized multivibrator frequency. This is shown in the drawing of Fig. 11-34. Synchronization is assumed to already exist in this drawing. Every third pulse (numbers *1*, *4*, *7*, *10*, and *13* in the drawing) causes the grid voltage to reach the cutoff bias of the tube prematurely and thereby start conduction. The conduction periods of the tube are indicated by that part of the grid-voltage waveform above the cutoff bias of the tube; below it,

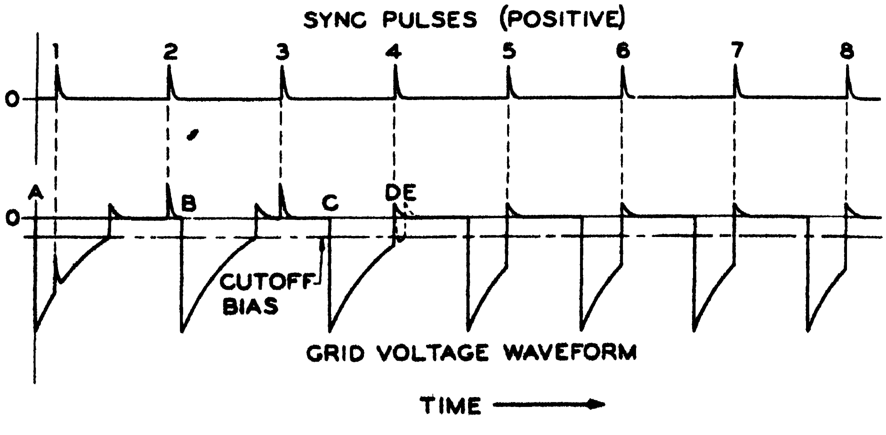

Fig. 11-32.—The grid waveform of a multivibrator being synchronized by positive pulses whose frequency is slightly greater than the multivibrators free-running frequency.

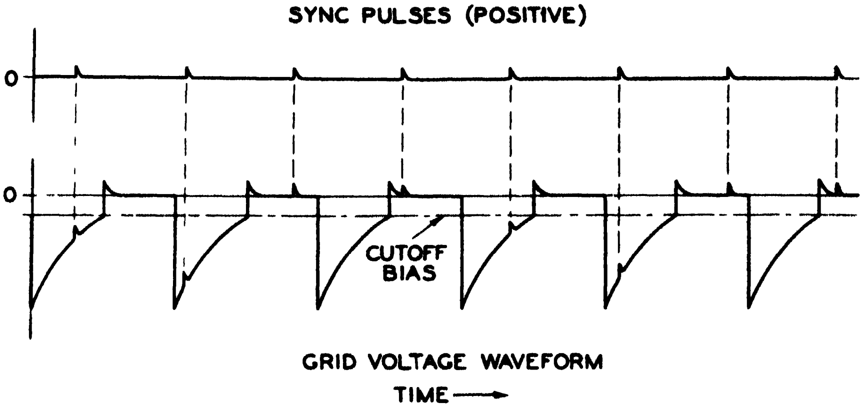

Fig. 11-33. — Insufficient pulse amplitude may not cause synchronization as indicated here.

the tube is nonconductive. Pulses *2, 5, 8,* and *11* are applied when the tube is in the conducting state and, therefore, have no synchronizing effect. Pulses *3, 6, 9,* and *12* are applied when the tube is nonconductive. Although these pulses reduce the bias on the tube, they are not strong enough to bring it to cutoff. Each group of pulses just discussed occurs at the same point in each cycle of synchronized multivibrator operation. For the case illustrated, the frequency of the sync signal is three times that of the synchronized multivibrator, that is, $f_s = 3f_o$.

Since the conductive and cutoff action of both tubes of a multivibrator are interrelated, then the synchronizing effect is coupled from one tube to the other and all waveforms line up in accordance with the synchronized grid waveform; synchronization of the circuit is, therefore, considered complete.

When a *negative* pulse is superimposed on the grid-voltage waveform, it will effect synchronization only when it is applied during the *conduction* period of the tube. When applied during the nonconduction period, it will not have any effect on locking-in the two signals.

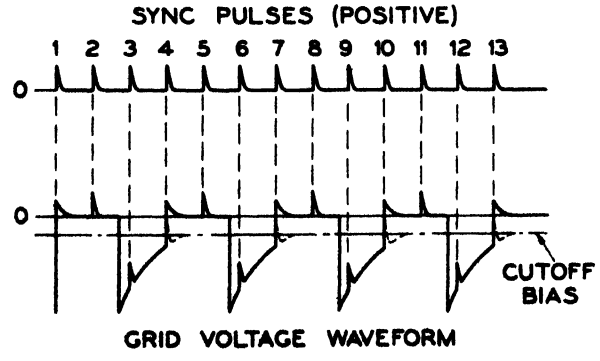

Fig. 11-34.—Synchronization of a multivibrator where the frequency of the sync pulses is three times the synchronized multivibrator frequency.

This is exactly the opposite of what happens when a positive pulse is applied. At first thought, one would then believe that the negative pulse, when applied to the grid must, during conduction periods of the tube, be strong enough to reduce the grid voltage of the tube to which it is applied from zero to a negative value great enough to drive the grid beyond cutoff. This is not necessary; let us see why.

In order to understand the phenomenon when a negative pulse is applied, we will study the grid-voltage waveforms of both tubes of the multivibrator. Refer to Fig. 11-35. Part (A) of this drawing indicates the negative pulse input; part (B) shows e_{g1}, the grid voltage of tube *V1*, to which the negative pulse is applied; part (C) is e_{g2}, the grid-voltage waveform of the other tube, *V2*. The important thing to remember about multivibrators in order to comprehend the analysis to follow is that the signals at the respective grids are 180° out of phase with each other.

The frequency of the grid-voltage waveform from time *a* to *e* is that of the free-running multivibrator, because during that time the sync pulses are not effective in initiating synchronization. The first negative pulse reduces grid voltage e_{g1}, but not by enough to reach the cutoff value of the tube. This pulse is, however, amplified by the tube *V1*, to which it is applied, and appears at the grid of tube *V2* as an *amplified positive pulse*. This is indicated at time *b*. The positive pulse on grid voltage e_{g2} is not great enough to raise the grid voltage to cutoff, therefore, the circuit continues to oscillate at its natural frequency.

Negative pulses *2* and *3* are superimposed on grid voltage e_{g1}, but occur during the nonconduction period of *V1*. The result is that the bias on the tube is made more negative at times *c* and *d*. Since tube *V1* is cut off during these moments, there is no coupling effect from tube *V1* to *V2*, and so the grid voltage of tube *V2* remains unchanged. The frequency and phase rela-

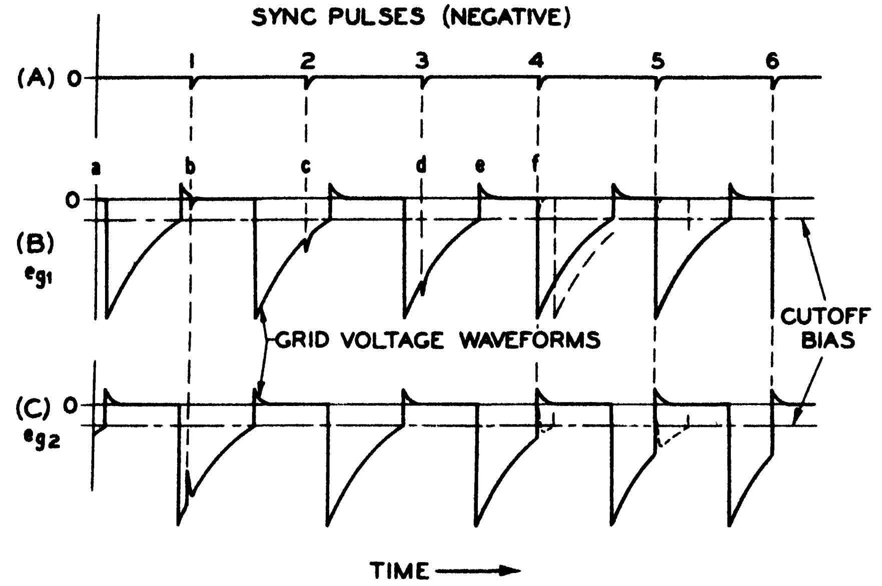

Fig. 11-35.—Graphical analysis of multivibrator synchronization by means of negative sync pulses.

tionship between the pulse signal and that of the free-running multivibrator are such that at time *f*, during pulse *4*, synchronization is initiated.

Negative pulse *4* reduces the e_{g1} bias, but is not great enough to reduce it to cutoff. This we already know from the action of pulse *1*. However, pulse *4* is applied to the grid at a later time, nearer the end of the conduction period of tube *V1*. When the negative pulse is coupled from *V1* to the grid of tube *V2*, it will be of the same positive amplitude as that indicated at time *b* on curve e_{g2}, because the amplification factor of tube *V1* remains the same. However, at time *f*, the value of e_{g2} is less negative than at time *b*. The amplitude of this positive pulse on the e_{g2} curve at time *f* is great enough to overcome the existing negative voltage to the point where the cutoff bias of the tube is reached. This means that tube *V2* will start conducting at an earlier time than usual, increasing its effective frequency of operation. The dotted parts of the grid-voltage curves after time *f* indicate what the continued shape of the curves would have been without synchronization.

After time *f*, each sync pulse will bring about a premature conduction of tube *V2*, and a few cycles after the initiation of synchronization, the frequency of the multivibrator will fall in with that of the sync pulses and synchronization will be complete. In the case under discussion, the frequency of the sync signal and that of the synchronized frequency of the multivibrator are equal, that is $f_s = f_o$. It is also possible to have f_s greater than f_o, similar to that indicated in Fig. 11-34.

The amplitude of the negative pulse does not have to be great enough to drive the tube to which it is applied to cutoff. It must be strong enough, however, when multiplied by the gain of this tube, to cause the second tube to start conducting prematurely.

Pulse Synchronization at Multiples of Sync Frequency

It is also possible to obtain synchronization when the frequency of the sync signal is less than that of the synchronized multivibrator at some *integral* ratio. When this occurs, only that particular cycle of oscillation being synchronized is controlled by the sync signal. For example, if the pulse repetition rate of a sync signal is 1,000 cps, and the multivibrator frequency is 2,000 cps, every second cycle of multivibrator oscillation will be synchronized, with the other cycles in-between the synchronized cycles "falling in line."

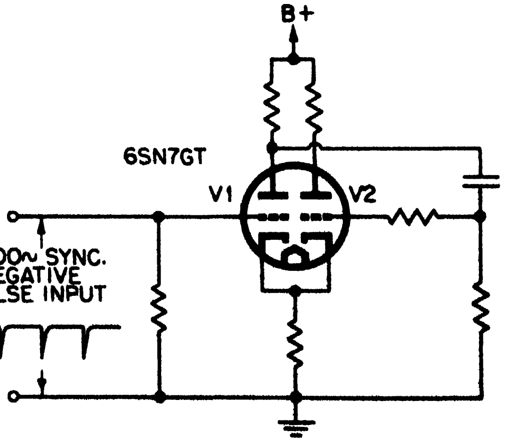

Fig. 11-36.—A typical cathode-coupled multivibrator circuit used in the experiment where the frequency of the negative sync pulses are less than that of the synchronized multivibrator oscillations.

On the basis of this type of synchronization, the fundamental frequency output from one rigidly controlled multivibrator can be made to control another relaxation oscillator or a group of oscillators. This is exactly what is done in the synthesizer discussed in Chapter 20. A standard sine-wave signal, obtained from a crystal-controlled oscillator or a tuning fork, is used to synchronize the fundamental multivibrator. The frequency of the sine-wave standard can be equal to or some integral *multiple* of the synchronized fundamental multivibrator frequency.

The controlled frequency of the fundamental multivibrator is 500 cps. The square-wave output from this circuit is differentiated into sharp pulses and then the positive pulses are clipped, leaving only negative pulses, which are then used to synchronize the multivibrator circuits operating on higher-order harmonics of 500 cps. Fig. 11-36 shows a typical circuit of any one of the multivibrators, indicating the negative pulse input. Note that it is a modified form of the cathode-coupled multivibrator.

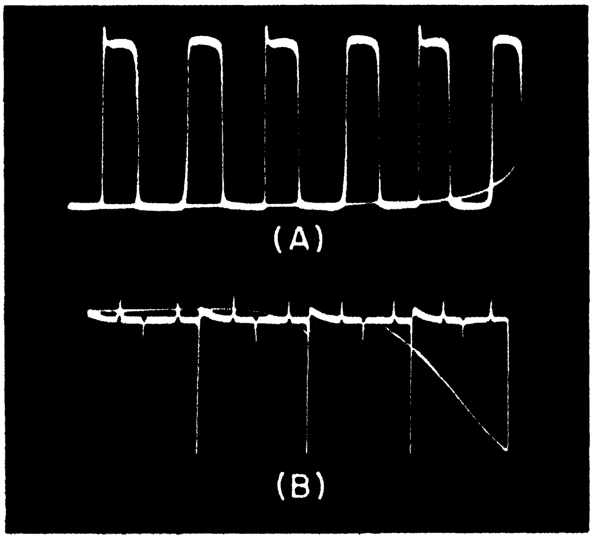

Fig. 11-37.—When the multivibrator is operating at twice the sync pulse frequency, the plate voltage waveform of (A) and the grid voltage waveform (B) results. Note the sync pulses appearing at every second cycle of (A).

The negative sync-pulse input is fed to the grid of *V1*. This means that the sync signal will appear in the plate circuit as a positive pulse. If the input pulse is strong enough, a picture of the square-wave output from the plate of *V1* will show small pulse pips on the leading edges of those square-wave cycles that are synchronized. In order to indicate this phenomenon, a number of oscillogram waveforms were taken at the plate of *V1*, for similar multivibrators operating at various harmonic frequencies of the 500-cps negative-pulse input.

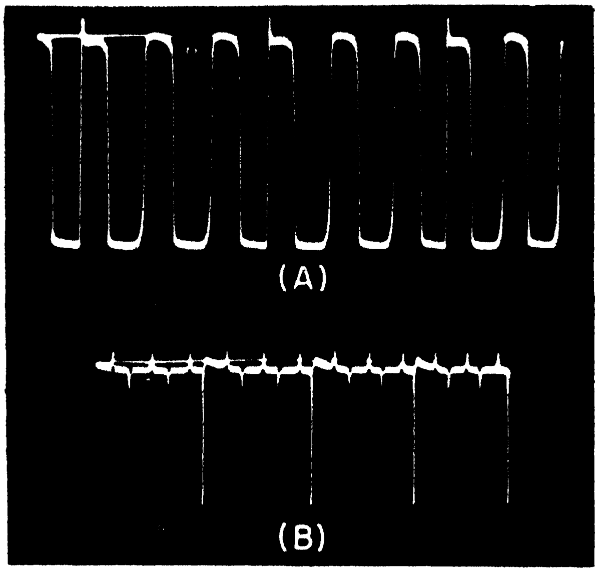

Fig. 11-38.—The plate voltage (A) and grid voltage (B) of the multivibrator when the sync pulse frequency is one-third that of the multivibrator frequency.

The photo at the *V1* plate of a 1,000-cps multivibrator is illustrated in part (A) of Fig. 11-37. Note the sync pulses are positive and appear on every other cycle of the square wave. This photo readily indicates that $f_o = 2f_s$. The oscillogram at part (B) of the same figure represents the grid voltage at pin *1* for the same 1,000-cps multivibrator. The long negative pulses represent the sync signal, which occurs at every other small negative pulse. The negative sync pulse is actually superimposed over a small negative pulse.

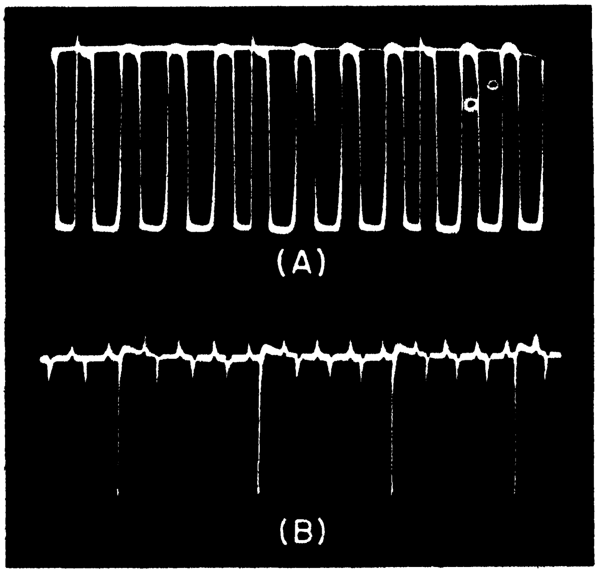

Fig. 11-39.—When the sync pulse frequency is equal to one-fourth the multivibrator frequency, the plate and grid voltages that result are indicated in parts (A) and (B) respectively.

Figs. 11-38A and B, respectively, show the plate and grid voltages from the circuit of the multivibrator in question when the frequency of the multivibrator is equal to 1,500 cps. Note that the sync pulses in each photo occur at every third cycle of the signal, indicating that $f_o = 3f_s$. When the multivibrator frequency is equal to 2,000 cps, the resultant plate- and grid-voltage waveforms will have the sync pulses appearing at every fourth cycle of the signal. This is indicated by the oscillograms of Fig. 11-39, in parts (A) and (B), respectively. For these pictures, $f_o = 4f_s$.

Synchronization of Blocking Oscillators

The method of synchronization for blocking oscillators is similar to that for multivibrators. Referring to Fig. 9-77 in Chapter 9, you will notice that the discharge part of the grid-voltage waveform of the blocking oscillator is similar to that of the multivibrator. It should be remembered, however, that a blocking oscillator normally consists of one tube. The synchronizing signal is usually injected into the grid circuit of the tube to cause the bias to increase at least up to the cutoff value and thereby make the tube conduct prematurely. Better control can be had by using pulses of short time duration.

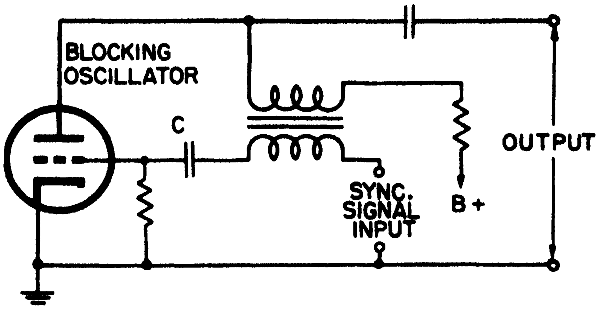

Fig. 11-40.—A typical method of introducing the sync signal to a blocking-oscillator circuit.

A typical method of introducing the sync signal to a blocking oscillator is illustrated in Fig. 11-40. In this circuit, the sync signal is coupled to the grid through the grid winding of the transformer and coupling capacitor *C*.

CHAPTER 12

PHASE AND FREQUENCY MEASUREMENTS

In many radio circuit problems in different fields of communication, it is often desired to know the phase difference between two voltages, two currents, or between a voltage and a current. In other cases, it is very important to determine the frequency of some particular signal. The oscilloscope is a very handy instrument for such phase and frequency measurements.

In the analysis of such types of measurements with the oscilloscope, we have to consider the application of an a-c signal to both sets of deflection plates. Consequently, in the process of studying the use of the oscilloscope for phase and frequency measurements, we will also be examining the action of the electron beam under the combined influence of both a-c signals.

A-C Voltages on Both Sets of Plates

When a voltage is applied to each set of deflection plates, the pattern which appears on the screen is the result of the continuous displacement of the spot in two dimensions at every instant of time and always acting at right angles to each other. The amplitude of these voltages, whether these voltages are alternating or direct, is going to influence the shape of the resultant pattern. In applying an a-c voltage to each set of deflection plates, another factor is very important in determining the shape of the resultant pattern on the screen. This factor is the *relative* phase relationship between the two a-c signals.

If the amplitudes of both a-c voltages are constant, but the phase of one or both is changed, such that the initial phase relation between them is no longer the same, then the pattern on the screen will also change. In like manner, if the initial phase relation is held constant and the amplitude of either one or both signals is changed, then the pattern on the screen will also change. Whatever the situation, the shape of the pattern is going to be determined by the phase relation between the instantaneous deflection-voltage fields, as well as by the amplitude relationship. Despite the fact that the phase relationship does influence the pattern, when this relation is unknown, the pattern nevertheless still conveys certain information.

It is also of interest to realize that simple as well as complex patterns may result from the application of a-c voltages to both plates. However, regardless of the nature of the pattern, it is important to understand that it is created by the movement of a *single* spot; each of the lines, no matter how many appear upon the screen, is traced by a single spot in motion, and the resultant pattern is due to the recurrent tracing of that path by the moving spot.

Any type of a-c voltage, applied to the two sets of plates, may be displayed on the screen. They can be sine waves or distorted waves, modulated waves or unmodulated waves. The only limitations imposed relate to the ability of the amplifiers within the oscillograph to pass or amplify the voltage at the frequencies of these waves. Simple patterns appear when simple waves are placed upon the respective plates, and when they bear some simple frequency relation to each other. If the nature of the voltages applied to both sets, or to one set of plates, is complex, due to distortion or modulation, the resultant pattern will show the complexity of the voltages. In Chapter 9, "Linear Time Bases," the basic graphical analysis of a-c voltages on both sets of deflection plates was already considered, where one voltage was a sawtooth waveform.

Lissajous Figures

When a point, or spot on the oscilloscope, undergoes two simple harmonic motions, which at any one instant of time are at right angles to each other, the resultant movement of this point or spot traces out a curve which is called a *Lissajous* figure.[1] Such patterns are very important because they are extremely useful in measurements of relative phase and frequency relations between two a-c signals. If the phase, frequency, or amplitude of either or both signals is changed, the resulting Lissajous figure will also change.

Phase measurements are usually made when the frequencies of the two input a-c signals are equal to each other. Therefore, Lissajous figures that result from a phase difference between two input signals of the same frequency are often referred to as phase-difference patterns. When the frequencies of the input signals are not a 1:1 ratio, Lissajous figures are generally used for frequency measurements.

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[1] Called thus after a French professor of the same name who first demonstrated such figures in 1855 by the use of beams of light and two plane mirrors at right angles to each other.

From this brief introduction to Lissajous figures, we see that they are used to measure both phase and frequency. In this chapter we will study various types of Lissajous figures under different conditions of phase and frequency. We will first analyze the resulting curves theoretically by graphical construction and then show some actual oscillographic waveforms. Phase measurements will be studied first, and afterward, various methods of frequency measurements will be considered. In most examples illustrated, the input a-c signals will be sinusoidal.

In order to understand how the spot traces out a pattern when an a-c voltage is applied to both sets of deflection plates, we will study the very simple situation where both a-c signals are in phase with each other, and of the same amplitude and frequency. In the examples to follow, only one cycle of each applied signal will be used in the illustrations.

Two In-Phase Sine Waves Equal in Frequency

The drawings of Fig. 12-1 are a graphical analysis, similar to that in Chapter 9, how the spot moves in relation to the two applied a-c voltages. Part (A) is the sine-wave input to the vertical-deflection plates; part (B) is the sine-wave input to the horizontal-deflection plates. Part (C) shows the spot displacement for the individual sine waves, when each signal is considered as acting alone. From parts (A) and (B) of the drawing, we can tell by the direction in which the signal is traveling (i.e., orientation of time axis) that they are in phase with each other.

The numbered spots along the vertical axis of part (C) correspond to those same numbers indicated on

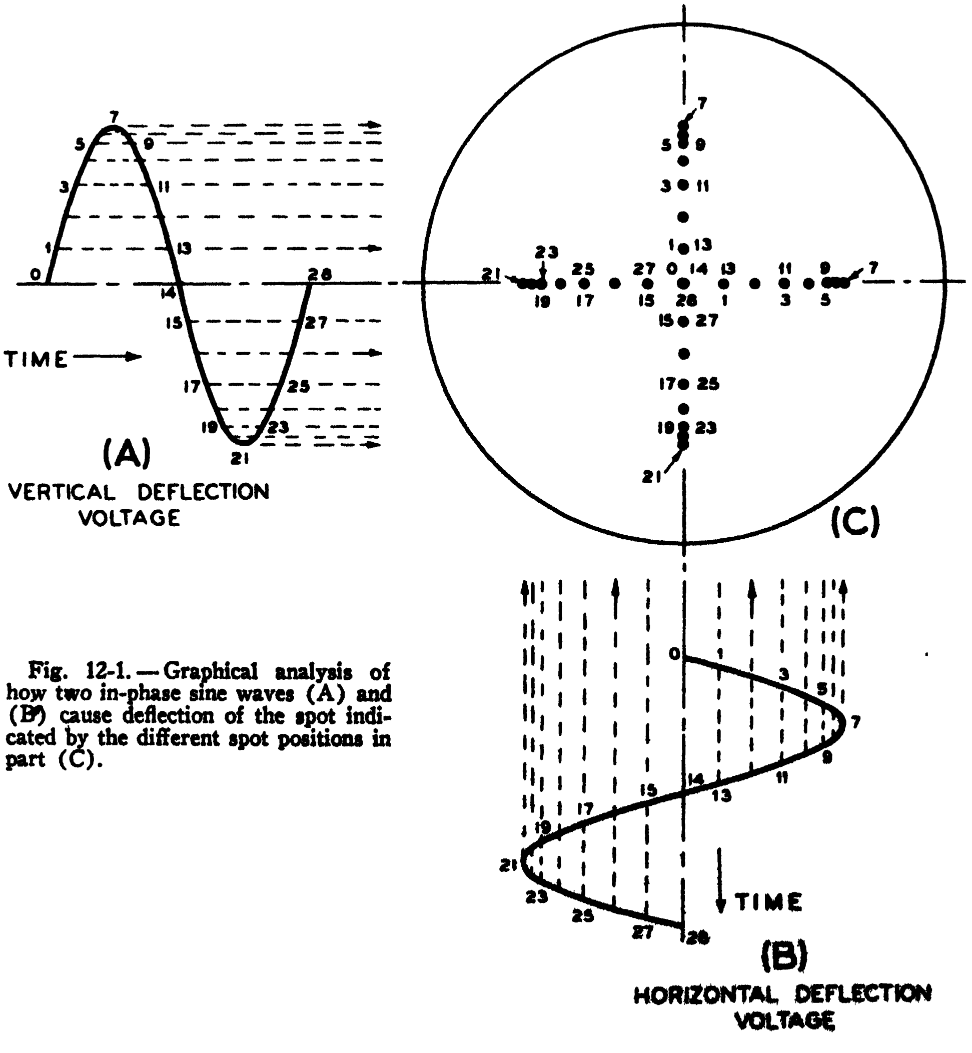

Fig. 12-1.—Graphical analysis of how two in-phase sine waves (A) and (B) cause deflection of the spot indicated by the different spot positions in part (C).

the sine wave at part (A). Similarly the numbered spots along the horizontal of part (C) correspond to those same numbers indicated on the sine wave at part (B). The direction of the dashed arrows from each sine wave lines up with their respective spot positions in part (C). Each spot of part (C), with the exception of those that occur during the zero and peak points of the sine waves, has two numbers indicating that the sine wave passes through that point twice during each cycle. The zero voltage spot has three numbers because each sine wave reaches zero voltage three times in each cycle. The peak points of each wave are, however, only represented by one number, because each sine wave passes through a single positive and negative peak during one cycle. The time interval between each point is uniform and the same for each alternating voltage. You will notice that, with the exception of the zero voltage markings, only odd numbers are indicated. The even numbers, although understood to exist, were omitted for the purpose of making the drawing clearer.

Before we analyze the formation of the actual pattern, let us for a moment consider the action of the spot when under the individual influence of either input signal. We will first consider the input to the vertical-deflection plates without any signal on the horizontal-deflection plates. Under this condition, a vertical line will appear on the screen. Let us see how this comes about by considering what happens during each quarter-cycle of vertical signal input.

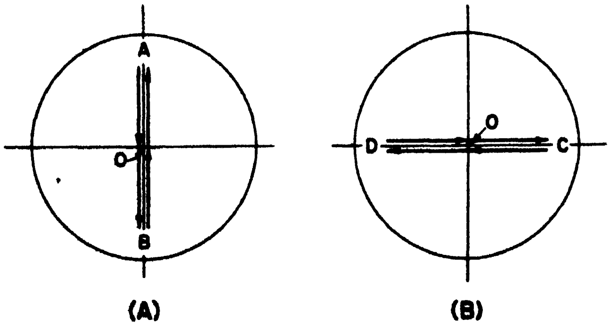

Fig. 12-2.—Simple vector diagrams illustrating the movement of the spot for one cycle of sine-wave input in quarter-cycle steps with that at (A) for vertical deflection and that at (B) for horizontal deflection.

Before any signal input is applied to the vertical-deflection plates, the spot remains stationary in the center of the screen. For the first quarter-cycle the voltage of the signal rises from zero to a positive maximum. This means that the spot will rise vertically from the center of the screen. This rise is indicated by line *OA* in Fig. 12-2A. For the next quarter-cycle of the sine wave, from 90° to 180°, the voltage decreases from positive maximum to zero. This means that the spot will fall from maximum back to the zero point, as indicated by the vertical line *AO* in Fig. 12-2A.

Although lines *OA* and *AO* in the drawing are shown as separate, actually the spot traverses the same points in rising to a maximum and then decreasing to zero. The two separate lines are drawn merely for the sake of illustration, but should be considered as being two coincident lines. This coincidence of lines should also be understood to exist in the rest of the analysis.

For the third quarter-cycle, from 180° to 270°, the sine-wave voltage decreases from zero to a negative maximum. The spot will, therefore, fall vertically from its center point and will trace out line *OB* as seen in Fig. 12-2A. For the remainder of the sine wave, the voltage increases from the negative maximum to zero again, which means that the spot rises once more but this time traces out line *BO* in Fig. 12-2A.

With a sine-wave input to the horizontal-deflection plates and with no signal on the vertical plates, only a horizontal line will be evident. (If the frequency of the sine wave is low enough, one half-cycle or less, one can actually see the movement of the spot tracing out the horizontal line.) The action of this signal on the horizontal-deflection plates for each quarter-cycle is similar to that just described for the vertical plates alone and is illustrated in Fig. 12-2B. Line *OC* is for the first quarter-cycle, line *CO* for the second quarter-cycle, line *OD* for the third quarter-cycle, and finally line *DO* for the last quarter-cycle. Actually, the line from *C* to *O* to *D* coincides with that from *D* to *O* to *C*.

In Fig. 12-3, we have illustrated what the final pattern will be when both input signals are applied simultaneously. Let us analyze this drawing in conjunction with those of Fig. 12-1 and see how this pattern is formed. Initially the spot remains in the center of the screen, indicated as point *0-0* in Fig. 12-3, and, after the passing of the first time interval, both the vertical and horizontal fields have a deflection of one unit. This positions the spot at *1-1* in Fig. 12-3. As time passes, the voltage of each wave increases, as they are in phase, and the resultant spot displacement is shown for each time interval with the corresponding numbers indicated as *3-3*, *5-5*, and *7-7*, in Fig. 12-3. Both waves now have reached their peak amplitude on the rising half of the first alternation. Now they both decrease in amplitude at a constant rate and the spot positions pass through the *same points as before*, since the time interval is the same.

In Fig. 12-3, these spots are indicated by numbers to correspond with those of Fig. 12-1, and are shown as *9-9*, *11-11*, *13-13*, and *14-14*, which is again zero.

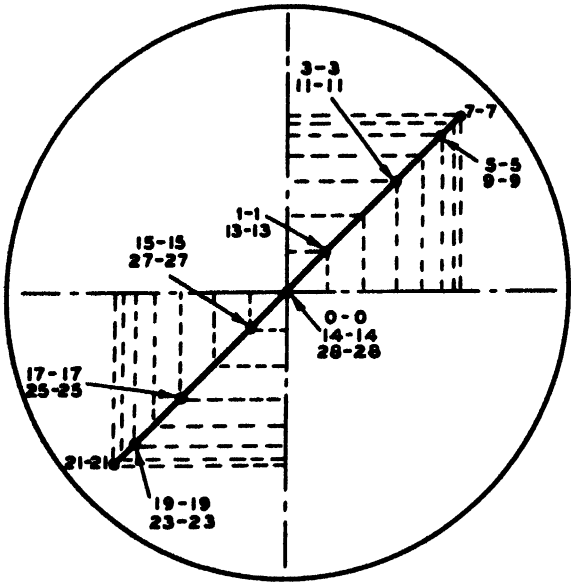

Fig. 12-3.—The simultaneous application of both deflection voltages produces the straight line shown here.

The positioning of the spots in this quadrant is due to the relative polarity of the plates as a result of the voltages applied.

In normal practice, the quadrant in the upper right-hand side of the drawing is referred to as quadrant I, and the others correspondingly become quadrants II, III, and IV in the *counterclockwise* direction from quadrant I. The horizontal axis to the right of the vertical line is considered positive and that to the left negative; also, the vertical axis above the horizontal line is positive and below it negative. Therefore, for the spot to fall in quadrant I, both instantaneous voltages must be positive. To fall in quadrant II, the voltage causing vertical displacement must be positive and that causing horizontal displacement must be negative. For quadrant III, both voltages must be negative, and for quadrant IV, the horizontal-deflection voltage must be positive and the vertical-deflection voltage negative. With this in mind, let us continue the analysis relative to the second half-cycle of the signal inputs.

As a result of the polarity of the respective vertical- and horizontal-deflection plates, the spot becomes positioned in quadrant III during this latter half-cycle of applied signals. The change in amplitude during the first quarter of the negative half-cycle causes individual displacements equal to amplitudes indicated by numbers *15, 17, 19,* and *21* in the vertical direction, and amplitudes indicated as *15, 17, 19,* and *21* in the horizontal direction (See Fig. 12-1C). The resultant spot positions are *15-15, 17-17, 19-19,* and *21-21* in Fig. 12-3. During the last half of the negative half-cycle, the vertical and horizontal amplitudes increase toward zero voltage, and they cause resultant displacement of the spot throuh positions *23-23, 25-25, 27-27,* and *28-28* in Fig. 12-3. As is evident, the result is a *diagonal* line of spots. Since these two voltages are actually continuous, instead of a number of spots, the result is an image which is a straight line.

A pattern of this type indicates certain definite conditions. First, the *frequency* of the two voltages applied to the vertical and horizontal plates is *identical,* and second, they are exactly *in phase* with each other. In other words, the instantaneous values of the two voltages are exactly alike with respect to time. A pattern of this type is independent of frequency. That is to say, this would be the pattern for any frequency, provided that both voltages were of the same frequency and exactly in phase. (Two voltages, two currents, or a voltage and a current are said to be in phase if they pass through their maximum and zero points in the same direction at the same time.)

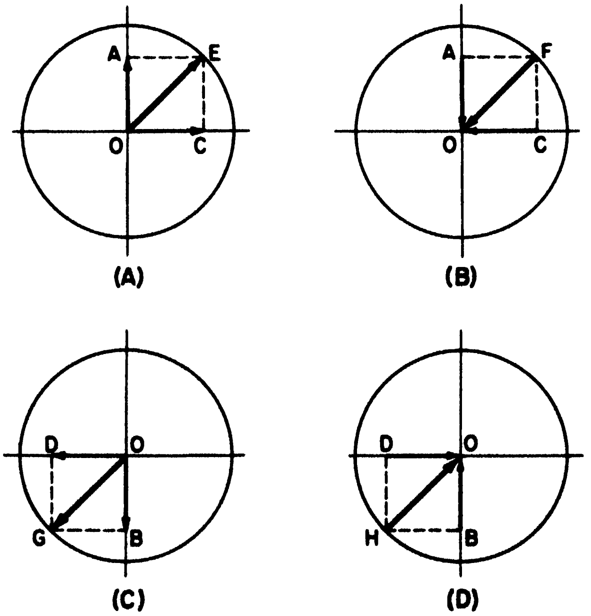

Fig. 12-4.—Vector diagrams illustrating how the straight-line image is formed; each drawing is for one quarter-cycle of signal input.

The formation of the straight-line image of Fig. 12-3 can also be demonstrated by simple plane vectors. As before, we will consider the action of the two input signals on the spot at quarter-cycle intervals. Refer to Fig. 12-4. In part (A), vectors *OA* and *OC* represent the individual spot displacements by the vertical- and horizontal-deflection voltages, respectively, during the first quarter-cycle. Vector *OE* is the resultant due to the addition of vectors *OA* and *OC*. This latter vector indicates the direction of spot travel (note arrow) which is away from the origin. For the second quarter-cycle, the individual vertical and horizontal spot displacements return to their original starting position. This is shown by respective vectors *AO* and *CO* in part (B), the combined action of which produces resultant vector *FO*. Note the direction of travel of this resultant vector and that it is coincident with the resultant vector of part (A), but in an opposite direction.

Vertical vector *OB* and horizontal vector *OD* in Fig. 12-4C indicate the individual spot positions for the third quarter-cycle (i.e., the beginning of the second half-cycle). Vector *OG* is the resultant of these latter two vectors and shows that the spot travels from the origin to point *G*, continuing in the same straight line established by the action on the spot during the previous half-cycles. For the final quarter-cycle, the spot traces back on the same path as for the third quarter-cycle. This is illustrated in part (D), where vectors *BO* and *DO* are the individual vertical and horizontal displacements and vector *HO* is the resultant of the addition of vectors *BO* and *DO*. Note that the direction of vector *HO* is headed *toward* the origin, point *O*, as compared to vector *OG* of part (C) which is directed *away* from the origin.

For the purpose of comparison, oscillographic waveforms duplicating the conditions of Figs. 12-1 and 12-3 were photographed and the photos rearranged to conform to the previous graphical illustrations. These photos are shown in Fig. 12-5 where (A) and (B) are the vertical- and horizontal-deflection voltages, respectively, and part (C) is the resultant pattern.

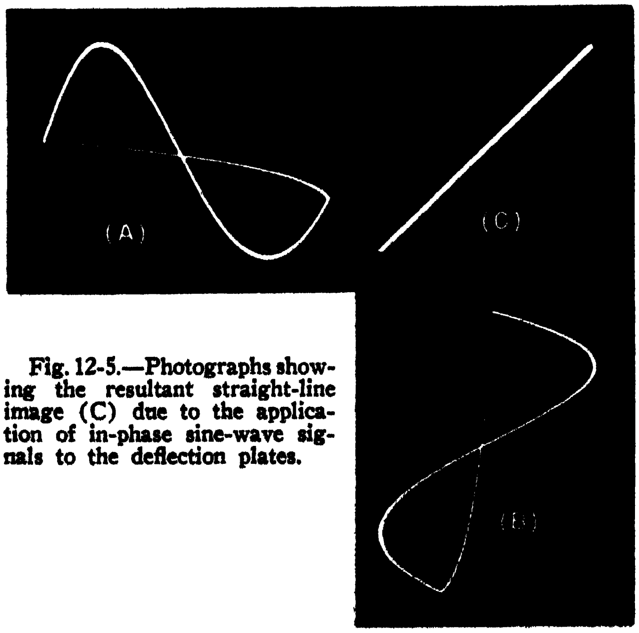

Fig. 12-5.—Photographs showing the resultant straight-line image (C) due to the application of in-phase sine-wave signals to the deflection plates.

The relationship between the axes (horizontal and vertical) and the resultant pattern, when the frequency and amplitudes of the two input signals are equal and in phase, can be photographically illustrated by a single picture. This photo is shown in Fig. 12-6A and is the result of a triple exposure. The photo was made in the following manner.

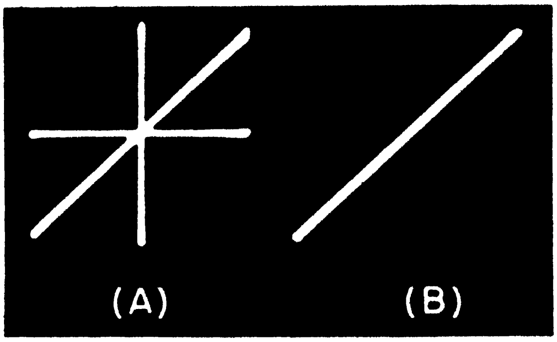

Fig. 12-6.—The straight-line images shown in relation to its horizontal and vertical axes at (A) and the image without the axes at (B).

With the sine waves applied to the vertical and horizontal input terminals of the oscilloscope, the gain of the vertical section is first reduced to zero; hence only the horizontal signal is effective and causes the horizontal or *X*-axis to appear on the screen. (The point where the horizontal gain control is left should be recorded.) This axis is then photographed and the film kept in place. Next the horizontal gain is reduced to zero and the vertical gain is increased, so that only the vertical or *Y*-axis appears on the screen. The vertical gain is increased to the point where the length of the vertical axis is made the same as that of the horizontal axis previously photographed. With this vertical axis established, the shutter of the camera is snapped once again, using the same film.

Thus far we see that, on one film, we have a double exposure illustrating the horizontal and vertical axes, and the same film remains as is for another exposure. For this third exposure, for the resultant pattern, the vertical gain control remains where it is and the horizontal gain control is increased to its previous position. With both input signals effective, the resultant phase-difference pattern appears upon the screen and is then photographed as the third exposure.

If the vector diagrams of Fig. 12-4 were superimposed over each other, the resultant diagram would be the same as Fig. 12-6A. Measurements of this latter figure will readily confirm the geometrical construction of the vector diagrams. The pattern in part (B) of Fig. 12-6 is the resultant phase-difference diagram without the axes. This photo was taken on the same film negative as Fig. 12-6A, but at a different part of the negative so that it would not interfere with that in part (A). Therefore, the single film negative was actually exposed four times.

In the remainder of this discussion on phase differences, a number of such patterns as Fig. 12-6A, including the horizontal and vertical axes, will be shown in conjunction with a resultant phase pattern. It should be remembered that these photos are triple exposures, all taken in the same manner as that just described.

From the analysis of Figs. 12-1 through 12-4 thus far, we find that when the two input signals are equal in frequency and amplitude, and when they are in phase with each other, the resultant image is a diagonal straight line, bisecting the angles of quadrants I and III. This *bisection* is obtained because the amplitudes of the two deflection voltages, and hence the deflection fields, are the same. Try to visualize the horizontal deflection reduced from that which is shown in the previous drawings, while the vertical amplitude remains constant. What would happen? Since the vertical amplitude would remain the same while the horizontal displacement would be less, the straight-line image would tend to approach the vertical reference line or Y-axis. This is proven by actual oscillograms as shown in Fig. 12-7, where part (A) indicates the resultant pattern including the axis and part (B) the pattern without the axis. Note that the angle between the vertical axis and resultant image is much smaller than the angle between the image and horizontal axis. If the horizontal deflection is continually decreased, the resultant pattern will gradually approach the vertical line until, when the horizontal deflection becomes zero, the image is a straight line along the vertical or Y-axis.

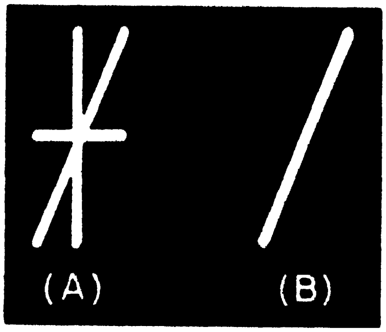

Fig. 12-7.—When the in-phase horizontal-deflection voltage is reduced, the resultant straight-line image with and without its axes appears in (A) and (B) respectively.

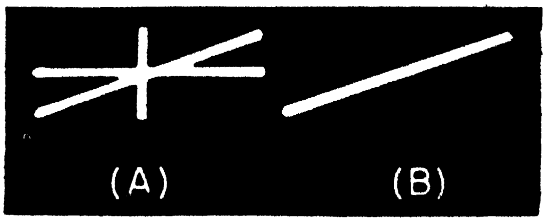

Fig. 12-8.—When the vertical-deflection voltage is reduced the patterns indicated here result.

If the reverse were true, that is, if the strength of the horizontal deflection were maintained constant and the vertical deflection were decreased, the straight-line image would tend to move toward the horizontal or X-axis, until, when the vertical deflection would be zero, the pattern would be a straight line along the horizontal axis. Examine the patterns of Fig. 12-8. These oscillograms illustrate a case where the sine wave causing vertical deflection is less than that causing horizontal deflection. That at (A) is the triple exposure, while that at (B) is the resultant pattern without the axes.

In practice, variation of the horizontal or vertical amplitude or gain controls of the oscilloscope rotates the pattern around the zero point. The exact position of the single-line pattern as it appears in quadrants I and III is not of great importance, for it indicates a zero phase difference as long as the pattern is a single line and is the result of the presence of both deflection fields. The exact angle that the pattern makes with respect to either the horizontal or vertical axis is also an indication of the relative amplitudes of the two voltages applied to the plates. The fact that the image is a single-line pattern lying in quadrants I and III indicates the zero phase relation between the two voltages.

Amplifier Considerations

What has been discussed thus far relative to the in-phase relationship between the two input signals for producing the straight-line image of Fig. 12-3 only holds if the signals are in phase *at the deflection plates* of the cathode-ray tube. It may be possible, however, to obtain this straight-line pattern under different conditions of signal input to the amplifiers of the oscilloscope. If the input signals to the amplifiers are in phase with each other and equal in frequency and amplitude, and if they have the same amount of vertical and horizontal amplification, as well as the same type of deflection output circuits, then the pattern of Fig. 12-3 will result. This is so because both input signals go through the same number of amplifying stages and, therefore, are subject to the same phase changes.

On the other hand, if the input signals to the two amplifiers are in phase with each other, but if there are an odd number of amplifying stages in the vertical amplifier and an even number in the horizontal amplifier, or vice versa, the vertical- and horizontal-deflection voltages at *the plates of the cathode-ray tube* will no longer be in phase with each other. In this particular instance, the signals at the deflection plates will be 180° out of phase with each other. If the input signals to the amplifiers, for this case of an odd-even number of vertical and horizontal amplifiers, were 180° out of phase, then the resultant signals at the *plates* of the cathode-ray tube would be *in phase*

The reason that the number of amplifying stages plays such an important role is that each stage inherently changes the phase between the input and output signals by 180°. In oscilloscopes, the amplifying stages are normally resistance-capacitance coupled, and introduce negligible phase shift.

The relative number of amplifying stages in each section is very important in the analysis of phase and frequency measurements. This should be remembered, especially for the remainder of this chapter. Whatever drawings are to follow, the final analysis of the pattern on the screen is to be based upon the action of the vertical- and horizontal-deflection voltages at the deflection plates of the cathode-ray tube.

Two Sine Waves 180° Out of Phase

Let us now consider the opposite situation where the signals at deflection plates of the cathode-ray tube are 180° out of phase with each other. For the case we are about to illustrate, we will change the starting time of the signal at the horizontal-deflection plates by 180° from what it was for the in-phase case of Fig. 12-1. The graphical illustration appears in Fig. 12-9. Part (A) is the signal on the vertical plates, part (B) that on the horizontal plates, and part (C) the spot displacement

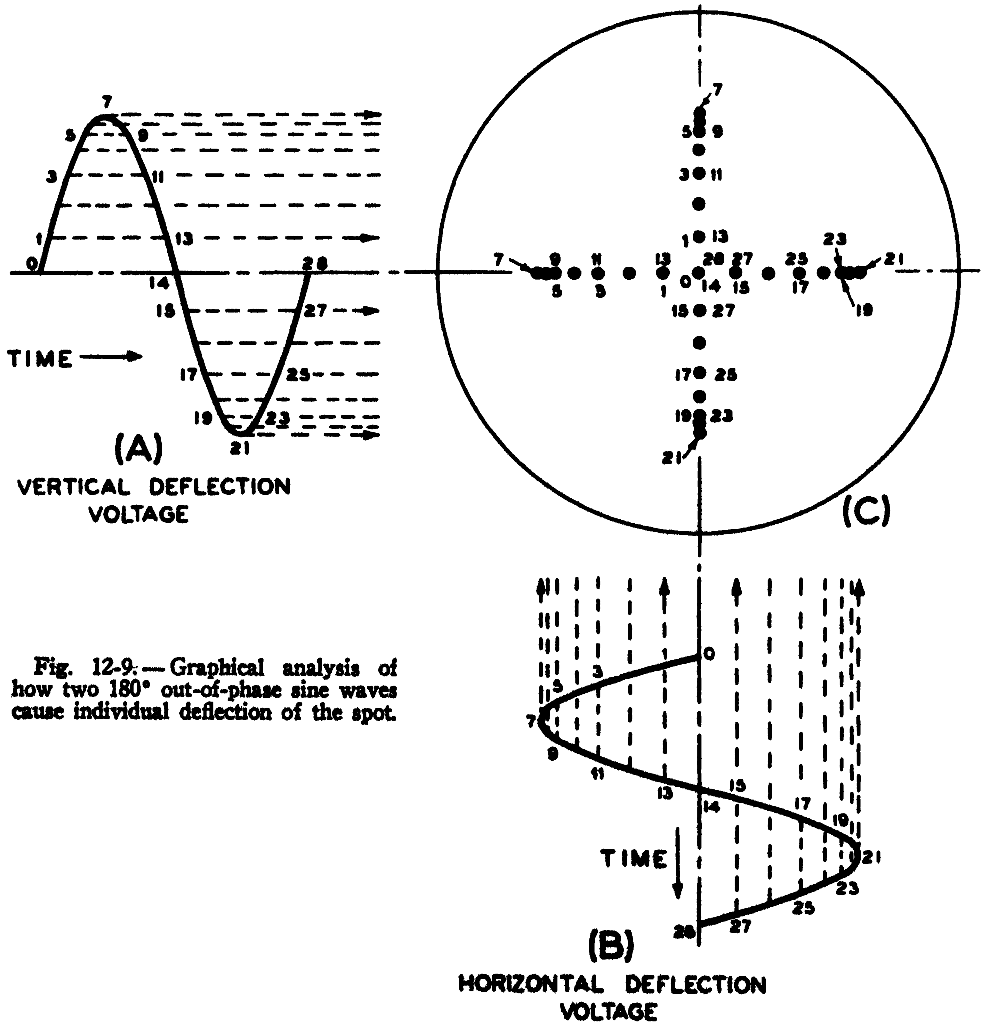

Fig. 12-9.—Graphical analysis of how two 180° out-of-phase sine waves cause individual deflection of the spot.

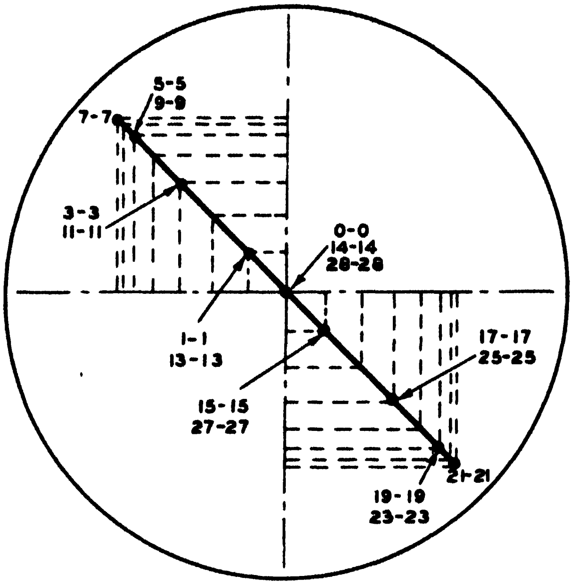

Fig. 12-10.—The simultaneous application of both deflection voltages of the latter figure produces the straight line as indicated.

for the individual sine waves when each is considered acting alone. The amplitude and frequency of these sine waves are equal, similar to the previous case. Both voltages are divided into the same equal time variations as before.

Both voltage start at zero, but increase in amplitude in opposite directions. With the individual spot positions as shown in Fig. 12-9C, the resultant pattern is as shown in Fig. 12-10. Note that the pattern now occupies space in quadrants II and IV. The geometrical construction of the pattern is along the same lines as discussed before. A similar pattern would be obtained if the horizontal voltage had been left intact as in Fig. 12-1B and the vertical wave of part (A) had been changed through 180°. A phase difference of exactly 180° is indicated only if the pattern is a single line, bisecting quadrants II and IV as shown. Additional details concerning the phase relation indicated by such patterns will be found later in the chapter.

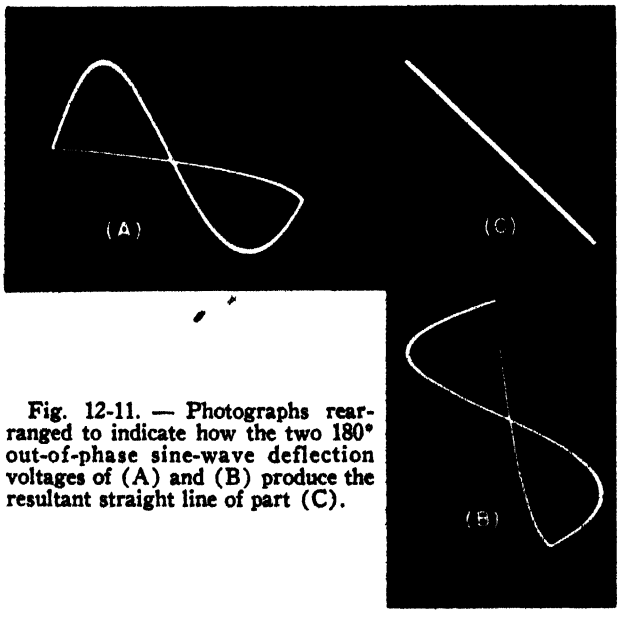

Fig. 12-11. — Photographs rearranged to indicate how the two 180° out-of-phase sine-wave deflection voltages of (A) and (B) produce the resultant straight line of part (C).

The waveforms of Figs. 12-9 and 12-10 have been duplicated on an oscilloscope and the resulting traces photographed. These photos have been arranged to conform to Figs. 12-9 and 12-10, and are illustrated in Fig. 12-11. Parts (A) and (B) are the sine-wave vertical- and horizontal-deflection voltages, respectively, and part (C) is the resultant straight-line image.

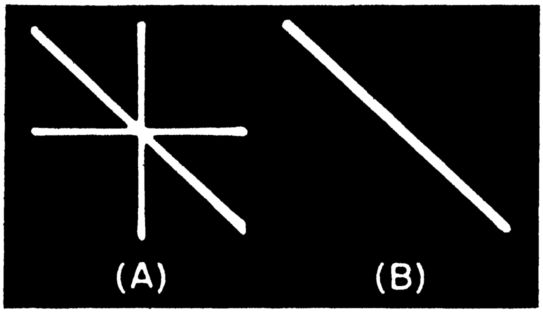

Fig. 12-12.—The straight-line image resulting from two sine waves 180° out of phase but of equal amplitude is shown at (A) with its axes and at (B) without its axes.

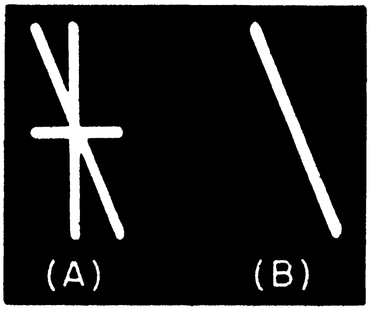

Fig. 12-13.—Similar to Fig. 12-12 except that the horizontal-deflection voltage is reduced.

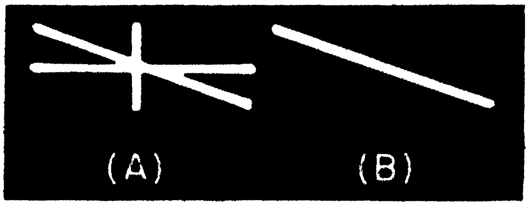

Fig. 12-13.—Similar to Fig. 12-12 except that the horizontal-deflection voltage is reduced.

Let us examine some oscillograms illustrating the 180° phase patterns for different amplitude relationships between the vertical and horizontal input signals. Figs. 12-12, 12-13, and 12-14, part (A) illustrate the pattern with both axes shown, while part (B) is the same resultant pattern without the axes. For Fig. 12-12, the amplitudes of both input signals are equal, which is evident from the diagonal line bisecting the angle formed by quadrants II and IV.

For Fig. 12-13, the amplitude of the horizontal input is less than that of the vertical input. This is immediately evident by comparing the lengths of the axes. Also note that the angle between the resultant pattern and the vertical or *Y*-axis, is smaller than that formed between the pattern and the horizontal or *X*-axis. The pattern that results when the vertical input signal is less than the horizontal appears in Fig. 12-14. Compare the lengths of the axes and the angles between the image and axes in this figure with those of Fig. 12-13.

The phase relation between two voltages applied to the vertical and horizontal plates, respectively, is not limited to 0° or 180°. *Any* phase difference between 0° and 180° can be recognized upon the screen by characteristic patterns. In connection with phase difference, while it is true that the complete cycle of angular rotation is 360°, certain phase-angle representations appear alike, when viewed upon the cathode-ray-tube screen. Phase-difference patterns representative of angles between 0° and 90° find their duplicates in patterns which vary in phase from 270° to 360°. Patterns which vary in phase from 90° to 180° find their duplicates in patterns which vary in phase from 180° to 270°. In other words, a pattern which indicates a 45° phase difference may also be 315°; a 30° phase difference may also be 330°; a 90° phase difference may also be 270°; and a phase difference of 135° may also be a 225° phase difference. For ordinary purposes, however, in connection with radio receiver, audio amplifier, and "ham" transmitter adjustments, phase-difference observations between 0° and 180° are entirely sufficient.

Two Sine Waves 90° Out of Phase

Let us examine the case where the two sine-wave signals are still equal in frequency and amplitude, but the phase relation between them is changed to 90°. Under this condition, a completely different type of pattern will result, namely, a circle. In the example to be illustrated, the sine wave causing vertical deflection

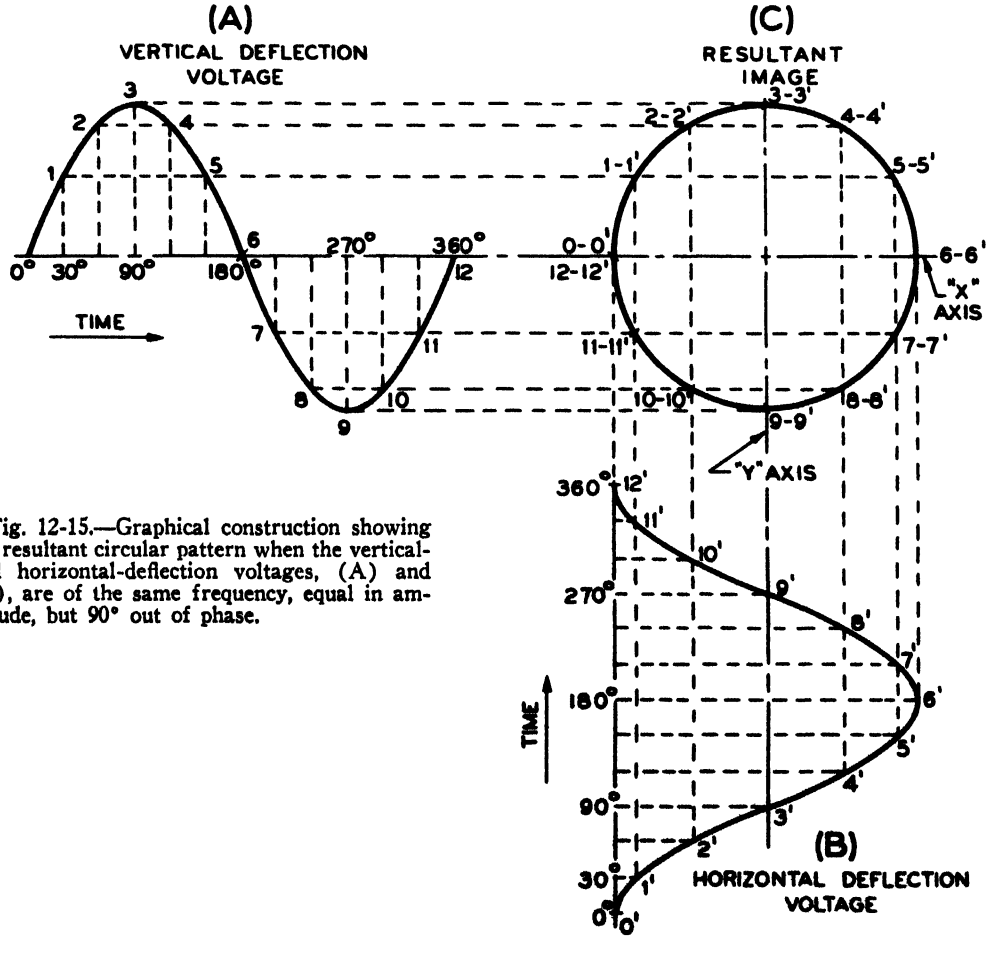

Fig. 12-15.—Graphical construction showing the resultant circular pattern when the vertical- and horizontal-deflection voltages, (A) and (B), are of the same frequency, equal in amplitude, but 90° out of phase.

is left, as it was previously, while that causing horizontal deflection is shifted in phase, so that both signals will be 90° (or 270°) out of phase.

The graphical construction of this 90° phase-difference pattern appears in Fig. 12-15. Parts (A) and (B) are the respective vertical- and horizontal-deflection voltages and part (C) is the resultant image, which is seen to be a circle. Note the phase difference between the sine waves at (A) and (B). The sine wave at (B) has its time axis going in the opposite direction than that normally shown, but for the case under illustration, this does not matter because this single cycle of signal is symmetrical about its 180° line.

You will note that the time reference line for the sine waves has been designated in degrees, with the actual divisions being at uniform time intervals of 30° each. This should make it easier to comprehend the resultant movement of the spot through a circle. At each 30° line of parts (A) and (B), perpendicular lines (shown dashed) are drawn until they touch the sine wave, as shown. Since there are 360° in one cycle and the cycle is divided into 30° intervals, there will be 12 dividing lines. These lines are thus numbered consecutively from 1 to 12 in the direction of travel. Those numbered from 1 to 12 for the wave at (B) have primes next to them to distinguish them from those of part (A), although similar numbers of each indicate that the same time lapse has occurred for each signal.

From these points of contact on each sine wave, other perpendicular lines are drawn until they intersect each other as indicated in the drawing. Those intersecting lines of the *same* number actually indicate the different spot positions due to the action of both deflection fields. With both waves starting out at their 0° time, we find that the spot traces out points *0-0′* to *12-12′* in consecutive order (in a clockwise direction). Points *0-0′* and *12-12′* are coincident.

The patterns involved in the geometrical construction of Fig. 12-15 have been duplicated on the oscilloscope. These oscillograms are shown in Fig. 12-16 in a similar arrangement to the line drawings of Fig. 12-15. Note how the peak-to-peak amplitudes of the sine waves at (A) and (B) are equal to each other and also to the diameter of the circle at part (C).

A circular pattern, therefore, denotes a case where the two input signals are *equal in frequency and in amplitude,* but have a phase difference of 90°. When the amplitudes of the two 90° out-of-phase input signals are not equal, however, the resultant pattern is an *ellipse.* Ellipsoidal patterns also result when the input signals are equal in frequency, but have a phase difference other than 0° or 180°. How, then, can we distinguish between an ellipse indicating a 90° phase difference and an ellipse of some other phase relationship? The answer lies in the way the ellipse orients itself with respect to the axes.

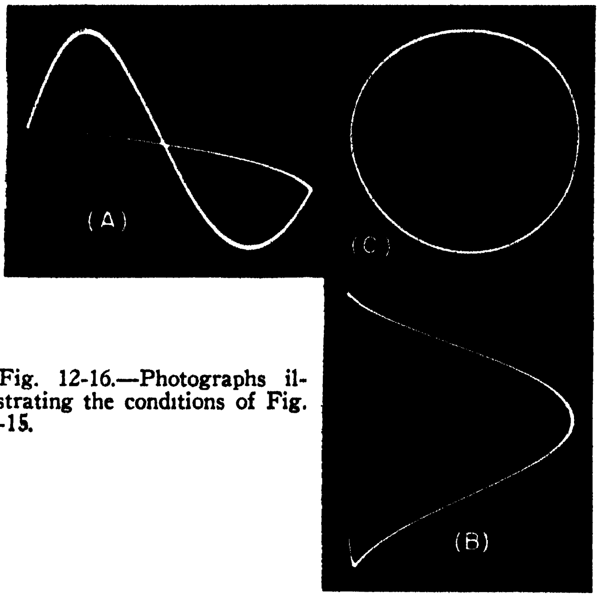

Fig. 12-16.—Photographs illustrating the conditions of Fig. 12-15.

Graphically speaking, the circular pattern is *symmetrical* about both the vertical or *Y*-axis, and the horizontal or *X*-axis. By this we mean that, if you could imagine folding the circle about its *Y*- or *X*-axis, the two halves of the pattern will coincide in each case. If only the amplitude relationship between two 90° out-of-phase input signals is changed, the ellipse that is formed will also be *symmetrical about both axes.* If the phase difference is other than 90° (or 270°), no matter what the amplitude relationship, the resulting ellipse will *not be* symmetrical about both axes.

If you will examine Fig. 12-15 once again, you will note that the spot travels in a clockwise direction to complete the circle. This is so because input signal (A) leads signal (B) by 90°, or (B) lags (A) by 90°. If either signal (A) or (B) had its relative phase position changed by 180°, both signals would represent a phase difference of 90°, but (B) would now lead (A) by 90°. A circular pattern would still result, but the spot would travel in a counterclockwise direction in tracing out the circle. We will say more about the direction of travel of the spot later on in this chapter.

The oscillographic waveforms of the following Figs. 12-17, 12-18, and 12-19 illustrate the resultant patterns of two 90° out-of-phase signals under different amplitude conditions. Of course, the frequency of each input signal is the same. As before, in each of the three figures, part (A) indicates the pattern with the axes

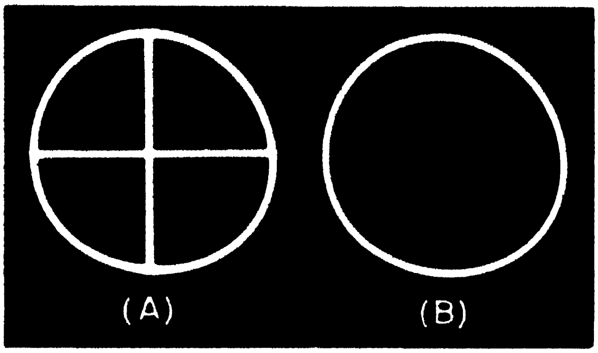

Fig. 12-17.—The circular pattern for the condition of Fig. 12-16 is shown with both axes at (A); at (B) it appears without the axes.

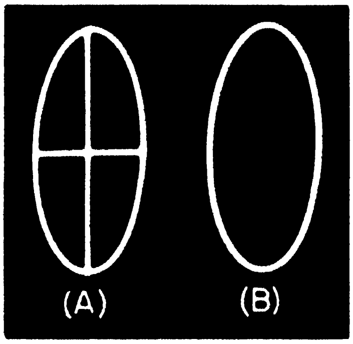

Fig. 12-18.—Where the horizontal-deflection voltage is decreased, the circle becomes an ellipse as indicated.

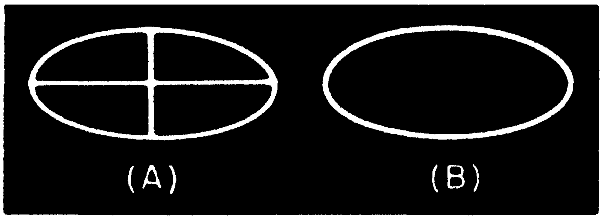

Fig. 12-19.—Where the signal voltage causing vertical deflection is decreased, the circle once again becomes an ellipse but in a different position, as indicated.

shown and part (B) without the axes. The circular patterns of Fig. 12-17 are for the case where the input signals are equal in amplitude. Note that both the *X*- and *Y*-axes of part (A) are equal. The patterns of Fig. 12-18 result when the amplitude of the horizontal-deflection voltage is less than that of the vertical-deflection voltage. Note that the length of the *X*-axis is smaller than that of the *Y*-axis. When the amplitude of the vertical-deflection signal is less than that of the horizontal-deflection signal, the pattern of Fig. 12-19 results. In this latter photo, the *X*-axis is larger than the *Y*-axis. Further examination of these last three figures readily indicates that the resultant patterns, the circle of Fig. 12-17, and the ellipses of Figs. 12-18 and 12-19, are symmetrical about both the horizontal and vertical axes.

Other Phase Patterns

Thus far we have seen what the resultant pattern is when the input signals, which are equal in frequency, have a phase difference of either 0°, 90°, 180°, 270°, or 360°. If the phase relationship between these two input signals is other than these values, the resulting pattern will be an ellipse in all instances. An elliptical pattern results regardless of the amplitude relationship between the two input signals. When the two signals have a phase difference anywhere between 0° and 90°, or between 270° and 360°, the major axis of the ellipse[2] will lie in quadrants I and III. If the phase difference is between 90° and 180° or between 180° and 270°, the major axis of the resulting ellipse will lie in quadrants II and IV.

Let us assume two input signals, one on the vertical plates and the other on the horizontal plates, such that they are equal in frequency and amplitude, and are in phase. What would happen to the resultant pattern if we were to keep the horizontal input signal as is, and just vary the phase of the vertical input signal from 0° to 360°? The pattern would vary from a straight line, bisecting quadrants I and III and would rotate in a counterclockwise direction going through ellipses of various degrees of eccentricity, circles, and straight lines, until it comes back to its original position.

The graphical analysis of Fig. 12-20 illustrates the above conditions. As in the previous diagrams, part (A) represents the vertical-deflection voltages, (B) the horizontal-deflection voltage, and (C) the resultant images. In this drawing, the horizontal voltage is kept constant and the vertical voltage has its phase varied in steps of 30° starting from the in-phase or zero-phase-difference position at *A1*. Sine waves at *A2, A3, A4, A5, A6,* and *A7* are shown leading sine wave (B) respectively by 30°, 60°, 90°, 120°, 150°, and 180°. From this graphical analysis, we can readily see the variation of the resultant pattern from *C1* to *C7*. Note that, as the pattern rotates from the ellipse at *C2* to that at *C3*, the ratio between the lengths of the major and minor axes of the ellipse decreases until, at pattern *C4*, the ratio becomes equal to unity as the pattern becomes a circle. The formation of the patterns is self-explanatory, since they are constructed along similar lines to the previous drawings.

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[2] In an ellipse, the major axis is the longer axis and extends from one extremity to the other and bisects the ellipse longitudinally. The minor axis, which is the smaller one, is perpendicular to and bisects the major axis. The major and minor axes of an ellipse are always mutually perpendicular.

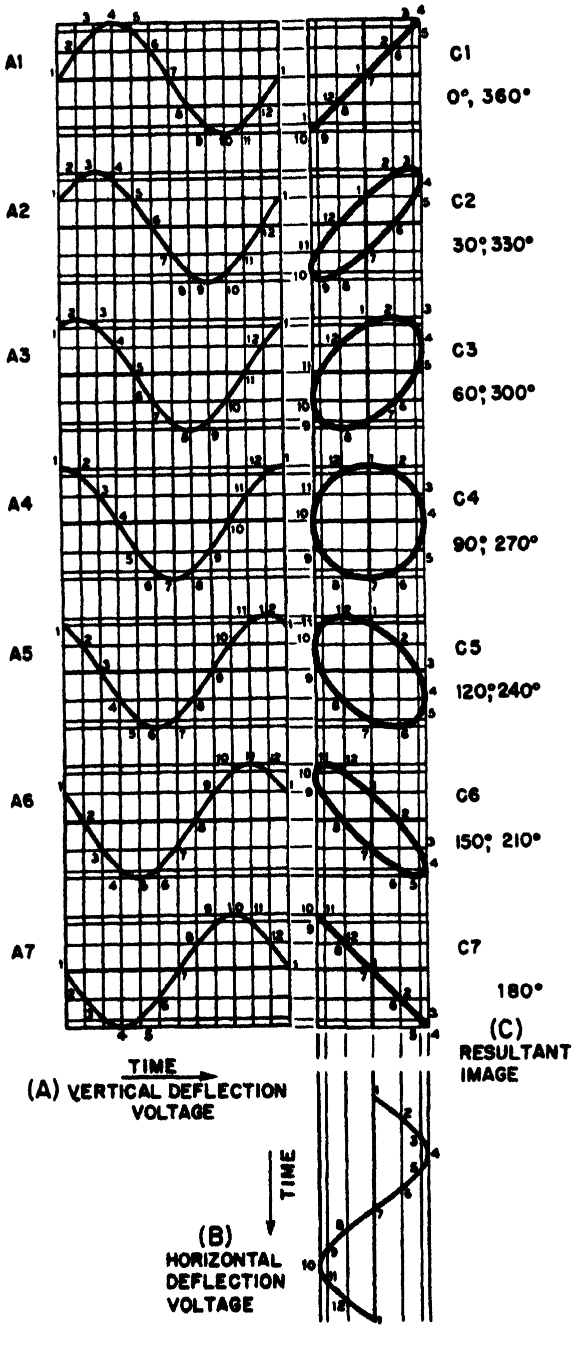


Fig. 12-20.—Where the vertical (A) and horizontal (B) input signals are equal in frequency and amplitude, the patterns that result in (C) for various phase conditions between 0° and 360° are indicated by *C1* to *C7*.

Next to each resultant image in part (C), you will see degree markings indicating the phase difference that such patterns represent. For instance, a pattern the shape of *C2* indicates either a 30° or 330° phase difference. If you examine the resultant patterns of *C2* through *C6*, you will note again that, in tracing each image, the spot travels in a *clockwise direction*. This occurs because the signals at *A2* through *A6* lead the signal at *B* by angles less than 180°. If the vertical-deflection voltage were advanced from its position at *A7* which represents a 180° phase difference, in steps of 30°, going through 210°, 240°, 270°, 300°, and 330°, the same patterns would be traced out once again. The resulting image at 210° phase difference will be the same as that for 150°, and so on, as indicated in the drawing; however, for those elliptical and circular patterns representing the 210°, 240°, 270°, 300°, and 330° phase differences, the spot traces out the image in a *counterclockwise direction*. This should be understood, although it is not indicated on the drawing of Fig. 12-20.

In the above analysis, we can say that the waves at part (A) lead the wave at (B) by angles of 0° through 360°. For those cases where the spot travels in a clockwise direction, which is from 0° to 180°, it is said that wave (A) *leads* wave (B) by a certain phase difference. In discussing phase differences relative to lead and lag, it is common procedure to discuss differences in terms less than 180°. Consequently for the situation of Fig. 12-20, where the waves at (A) are advanced beyond the 180° phase-difference point, that is, where the patterns are traced out in a counterclockwise direction, we say that wave (B) leads wave (A) by 30°, 60°, 90°, 120°, and 150°, instead of saying wave (A) leads wave (B) by 210°, 240°, 270°, 300°, and 330°, respectively.

In Fig. 12-21, line drawings of phase diagrams (input signals equal in frequency and amplitude) are drawn at intervals of 22.5° in order to show how the patterns vary at smaller phase differences. The degree values underneath each pattern readily indicate the possible phase differences represented by such patterns.

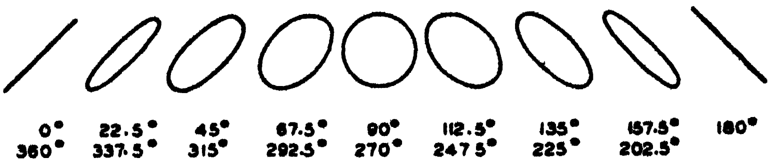


Fig. 12-21.—Different resultant patterns when the phase difference between the two input signals varies in steps of 22.5°.

Oscillograms of Elliptical Patterns

Now that we have graphically analyzed how the resultant elliptical pattern is obtained when the phase difference is other than 0°, 90°, 180°, or 270°, let us study a number of oscillographic waveforms of such elliptical patterns. In the following elliptical photos, Figs. 12-22 through 12-27, the phase difference is in steps of 45°. In part (A) of each drawing, the resultant pattern is indicated in conjunction with the vertical and horizontal axes, and for part (B), the resultant image appears without the axes. The axes serve as an ideal means of telling the amplitude relationship of the two input signals.

The elliptical patterns of Fig. 12-22 indicate a phase difference of about 45° (or 315°). Note that the X- and Y-axes are equal to each other which indicates that the horizontal and vertical signal inputs have the *same* amplitude. The elliptical pattern for this particular phase difference has its major axis *bisecting* quadrants I and III and its minor axis *bisecting* quadrants II and IV.

If the amplitudes are no longer equal, a differently shaped ellipse will result, even though the phase difference remains the same. The ellipses of Fig. 12-23 also represent a phase difference of 45° (or 315°), but the amplitude of the horizontal input voltage is now reduced from what it was in Fig. 12-22. The X- and Y-axes of Fig. 12-23A readily indicate the amplitude relationship. This elliptical pattern tells us that with the phase remaining equal and the horizontal-deflection voltage decreased, the major axis of the ellipse is

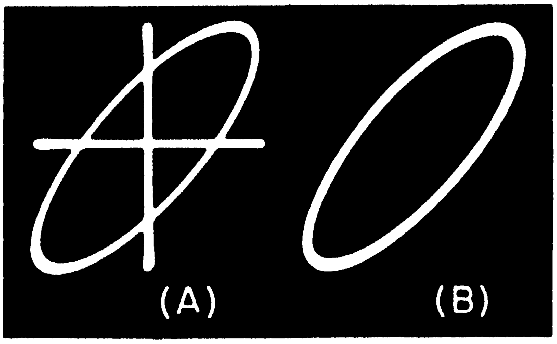


Fig. 12-22.—When the two input signals are equal in frequency and amplitude and have a 45° or 315° phase difference, the elliptical patterns shown here result.

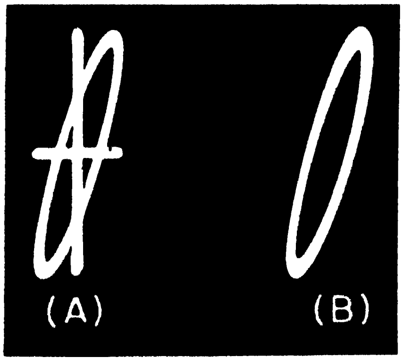


Fig. 12-23.—The same phase difference as in Fig. 12-22 but the axes of part (A) indicate that the horizontal input signal is decreased in amplitude.

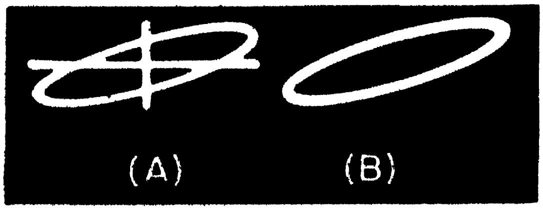


Fig. 12-24.—When the vertical-deflection voltage is decreased the ellipse of 45° or 315° phase difference will take on the position indicated here.

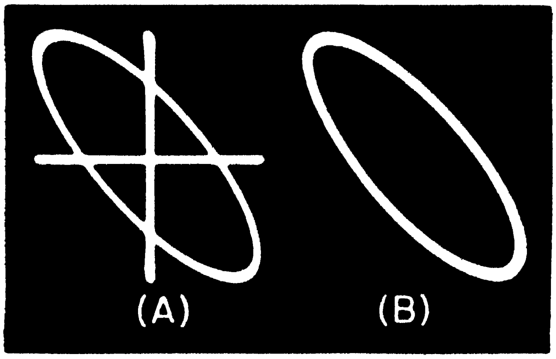


Fig. 12-25.—At a phase difference of 135° or 225° the two input signals, which are equal in frequency and amplitude, produce the resultant ellipse as indicated.

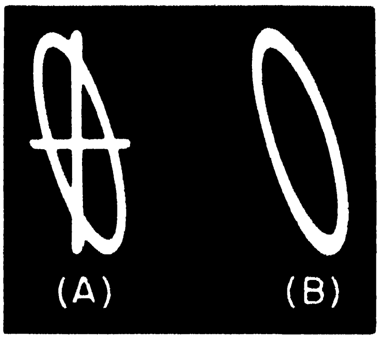


Fig. 12-26.—When only the horizontal-deflection voltage of Fig. 12-25 is reduced, the patterns shown here results.

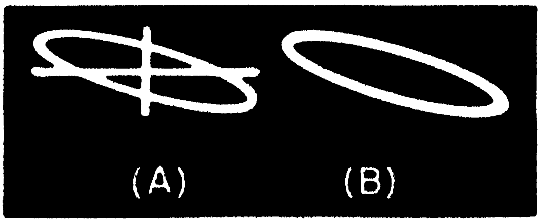


Fig. 12-27.—When the vertical-deflection voltage of Fig. 12-25 is decreased, the patterns shown here result.

rotated toward the Y-axis or vertical reference line of the co-ordinate system, but still *remains* in quadrants I and III.

If the vertical-deflection voltage were reduced instead of the horizontal voltage, the phase difference still remaining the same, the resulting elliptical pattern would have its major axis rotated toward the horizontal reference line on the X-axis. This is indicated in Fig. 12-24.

Let us now make the amplitudes of the two input signals equal again, but increase their phase difference by about 90° from the former case, which makes the phase relation about 135° (or 225°). When this is done, the resulting elliptical pattern will be as shown in Fig. 12-25. For this photo the major axis of the ellipse bisects quadrants II and IV while the minor axis bisects quadrants I and III. This is just the opposite of Fig. 12-22. When the horizontal-deflection voltage of Fig. 12-25 is reduced, the major axis of the resulting ellipse will rotate toward the Y-axis, as indicated in Fig. 12-26. This is a similar situation to that of Fig. 12-24, but the major axis of the ellipse of Fig. 12-26 is rotated in a clockwise direction toward the Y-axis, whereas for Fig. 12-24, it is rotated in a counterclockwise direction.

When the vertical-deflection voltage is reduced instead of the horizontal voltage, the resultant ellipse takes on the shape illustrated by the photo of Fig. 12-27. The phase difference is still 135° (or 225°). The major axis of the ellipse is oriented toward the X-axis and is rotated in the counterclockwise direction in reaching the position.

Calculating the Phase Difference from the Pattern

We have already shown that, according to the orientation and shape of 1 :1 ratio patterns, a fair idea of the phase difference can be obtained. There is a method, however, requiring a simple measurement on the pattern, that will enable us to calculate fairly well what the phase difference is. This method is indicated in Fig. 12-28 for an ellipse that portrays a phase difference of 45° (or 315°).

From this figure, you will note that two distances along the $+Y$-axis are marked off. Distance A represents that part of the curve where the ellipse *intercepts* the Y-axis. By drawing a line parallel to the X-axis and tangent to the top of the ellipse, the distance B is determined by the Y-distance at the place where this parallel line crosses the Y-axis. No matter what the plates and the equations in x represent the input to the tance B can never be less than A.

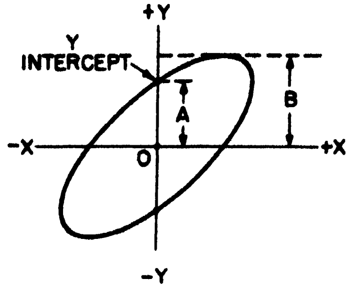


Fig. 12-28.—Method of calculating the phase difference from the pattern.

The phase difference from such patterns is calculated by the following formula

$$\sin \theta = \frac{A}{B} \tag{12-1}$$

where θ is the angle representing the phase difference and A and B are as indicated in Fig. 12-28. From this figure, we find that the ratio of A/B is approximately equal to 0.7, and the angle whose sine is equal to this value is found to be about 45°.

Although Fig. 12-28 shows the dimensions taken from the $+Y$-axis, the same measurement of A and B would result if they were taken from the $-Y$-axis. For the case under discussion, the amplitudes of the two input signals are equal, so that the distances from the origin point O to each $\pm X$ and $\pm Y$ intercept are equal. In like manner any tangent line will produce the same distance along any axis to which it is measured. Therefore, similar measurements made along either the positive or negative X-axis can also be used to determine the phase difference.

If the amplitudes of the two input signals are not equal, *either* the Y- or X-axis can still be used as the measuring reference. This is so because the *ratio* of A/B will always remain the same for a particular phase difference, regardless of the amplitude relationship of the input signals. For any one particular setting of amplitudes, the positive and negative Y-axes have the same A and also the same B measurement; this also holds true for measurements along the X-axis, as indicated in Fig. 12-29.

The ellipse in each part of this drawing is exactly the same, each indicating a phase difference of 45° or 315°, and the amplitude of the horizontal-deflection voltage is less than that of the vertical-deflection voltage. Part (A) indicates measurements along the Y-axis, whereas part (B) indicates measurements along the X-axis. For part (A), measurement $A1$ equals $A2$ and $B1$ equals $B2$; therefore, $A1/B1 = A2/B2$. For part (B), measurement $A3$ equals $A4$ and $B3$ equals $B4$; hence $A3/B3 = A4/B4$. Even though $A1$ and $A2$ are not equal to $A3$ and $A4$, and $B1$ and $B2$ are not

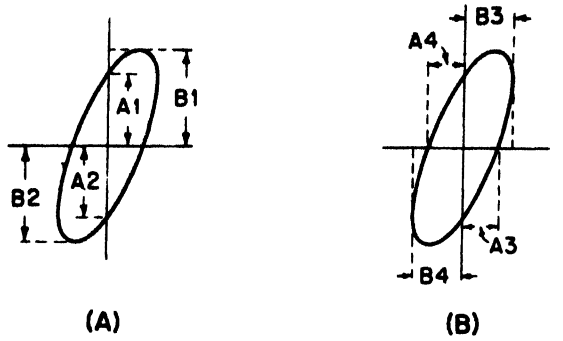


Fig. 12-29.—Similar method of calculating the phase difference from the pattern when the input signals are not equal in amplitude.

equal to $B3$ and $B4$, the ratios of $A1/B1$, $A2/B2$, $A3/B3$, and $A4/B4$ will all be the same. Consequently, we see that either the positive or negative half of the X- or Y-axis can be used to determine the phase difference, irrespective of the amplitude relationships of the two input signals.

For the pattern of Fig. 12-28 to represent a 45° phase difference, the spot has to trace out the ellipse in a clockwise direction. If the spot travels in the counterclockwise direction, the phase difference will be 315°. The direction of spot travel and the orientation of the major axis of the ellipse are two factors that should be known if the exact phase difference is to be calculated. Since we already know from the relative location of the patterns in the quadrants what the phase-difference limits may be, it is, therefore, a simple matter to calculate the phase difference from any elliptical pattern.

In Fig. 12-30 are shown seven different patterns, similar to Fig. 12-28, indicating phase differences from 0° to 300°, in steps of 30° each. From 0° to 180°, the spot traces each pattern in a clockwise direction, whereas for 180° to 360° each respective pattern is traced in a counterclockwise direction. In the above patterns, the phase angles are always expressed as angles of *lead* of the vertical-deflection voltage with respect to the horizontal-deflection voltage; i.e., a phase angle of 300° means that the vertical signal is *leading* the horizontal by 300°, and the pattern will be traced in a counterclockwise direction. Although the figure shows that, for each case, the phase difference is obtained from measurements along the positive Y-axis, any other axis could be used and the same phase angle would result For the sake of illustration, the amplitudes of the two input signals for all the patterns of Fig. 12-30 were assumed to be equal.

You will note that the ratio of A/B for parts (A) and (G) of Fig. 12-30 is equal to zero. The angle whose sine is equal to zero is either 0° or 180°. However, we have previously indicated that when the slope of the straight line is positive (i.e., the line lies in quadrants I and III) the phase difference is 0°; therefore, the phase differences for Figs. 12-30 (A) and (G) are as indicated. You will also note that the A/B ratio for parts (B) and (F) are both equal to 0.5. Since many sine tables only list angles up to 90°, from these tables, the angle whose sine is equal to 0.5 is indicated as 30°. Actually, the angle whose sine is equal to 0.5 can be either 30° or 150°. From our previous knowledge of the phase-difference limits with regard to the orientation of the ellipse, the phase differences would be as indicated in the drawing. A similar analysis holds true for the patterns of parts (C) and (E) of Fig. 12-30.

If the spot traces out the pattern in the counterclockwise direction, the calculation of the phase angle is a little different. Based on this assumption and on the knowledge of the orientation of the pattern, as

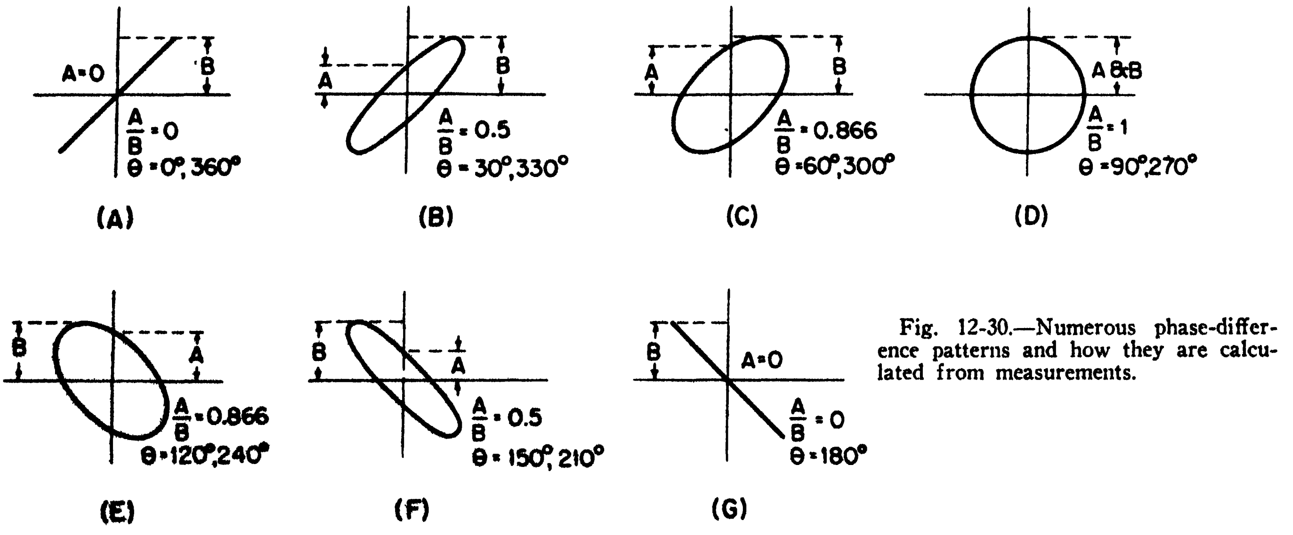


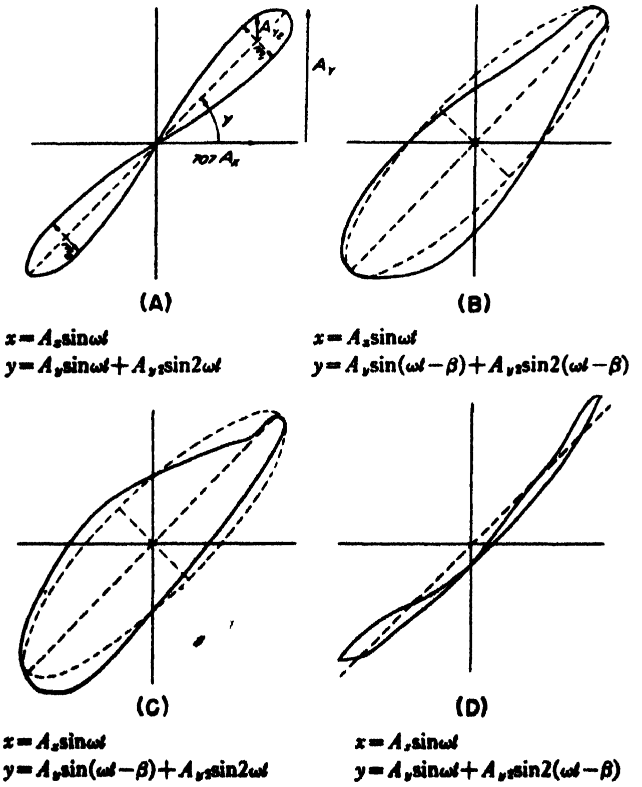
Fig. 12-30.—Numerous phase-difference patterns and how they are calculated from measurements.

shown in Fig. 12-30, the phase difference is expressed as follows

$$\theta = 360^\circ - \sin^{-1}\frac{A}{B} \qquad (12\text{-}2)$$

where the expression $\sin^{-1}$ means "the angle whose sine is." In other words, we are subtracting the answer of equation (12-1) ($\sin\theta = A/B$, which can also be expressed as $\theta = \sin^{-1} A/B$) from 360°. The validity of equation (12-2) can easily be checked from the A/B measurements of Fig. 12-30 by substituting them in the formula. For phase calculations from such patterns, it is best to have the image sharply focused (i.e., thin lines) and to use a finely divided co-ordinate scale for measurements.

The direction of travel of the spot can be rapidly checked by changing the phase relationship of the two input signals and noting the direction in which the pattern changes. The direction of the additional phase change must, of course, be known. If the pattern originally takes on the shape as illustrated in Fig. 12-30C, we do not know whether it represents a 60° or 300° phase difference, unless we also know the direction of spot travel. If we change the phase relationship in a known direction, say by adding an additional 30° change, a circle will result if the original phase difference is 60° (30° + 60° = 90°) or an ellipse smaller in area will be produced if the original phase difference is 300°, since now the resultant angle is 30° + 300° = 330°. *In other words, we are actually observing whether the minor axis increases or decreases.* In our problem, if it increases, the spot is traveling clockwise and the phase difference is 60°, and if it decreases, the spot is traveling counterclockwise and the phase difference is 330°.



(A)
$x = A_x \sin\omega t$
$y = A_y \sin\omega t + A_{y2} \sin 2\omega t$

(B)
$x = A_x \sin\omega t$
$y = A_y \sin(\omega t - \beta) + A_{y2} \sin 2(\omega t - \beta)$

(C)
$x = A_x \sin\omega t$
$y = A_y \sin(\omega t - \beta) + A_{y2} \sin 2\omega t$

(D)
$x = A_x \sin\omega t$
$y = A_y \sin\omega t + A_{y2} \sin 2(\omega t - \beta)$

Courtesy Rev. Sci. Instr.

Fig. 12-31.—Different types of patterns that result when the horizontal-deflection voltage is a sine wave and the vertical-deflection voltage a complex wave containing only a second harmonic and whose fundamental frequency is equal to the horizontal signal.

Other 1:1 Patterns

What has been discussed thus far relative to phase measurements, with the input voltages equal in frequency, was only based upon sine-wave signals. When either one or both of the input signals are not pure sine waves, the resulting patterns differ from straight lines, circles, or true ellipses. Let us first study a series of patterns where the input signal to the horizontal-deflection plates is a sine wave, but where the input to the vertical-deflection plates is a complex wave containing harmonics.

In Fig. 12-31[8] appear four different patterns that result when the input signal to the Y- or vertical-deflection plates also contains a second harmonic in addition to the fundamental. The makeup of the input signals is indicated underneath each curve. The equations in y are for the input to the vertical-deflection plates and the equations in x represent the input to the horizontal-deflection plates. From these equations, the general phase relationship between the x input signal and the components of the y input signal can readily be seen. The dashed lines indicate what the shape of the patterns would be for sine-wave input signals which are equal in frequency and amplitude.

For the pattern in part (A) of Fig. 12-31, both the fundamental and second harmonic of the vertical-deflection voltage are in phase with the horizontal-deflection voltage. The percentage of the second-harmonic amplitude compared to that of the fundamental amplitude of the y input signal, which is $A_{y2}/A_y \times 100$ can be calculated readily from the figure by measurement of A_y and m, the maximum opening of the pattern. We obtain then

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[8] Wood, L. A., "Phase measurements with the cathode-ray oscillograph," *Rev. Sci. Instr.*, vol. 2, pp. 644-648, October, 1931; Figs. 12-31 and 12-32, and associated equations were reprinted from this article.

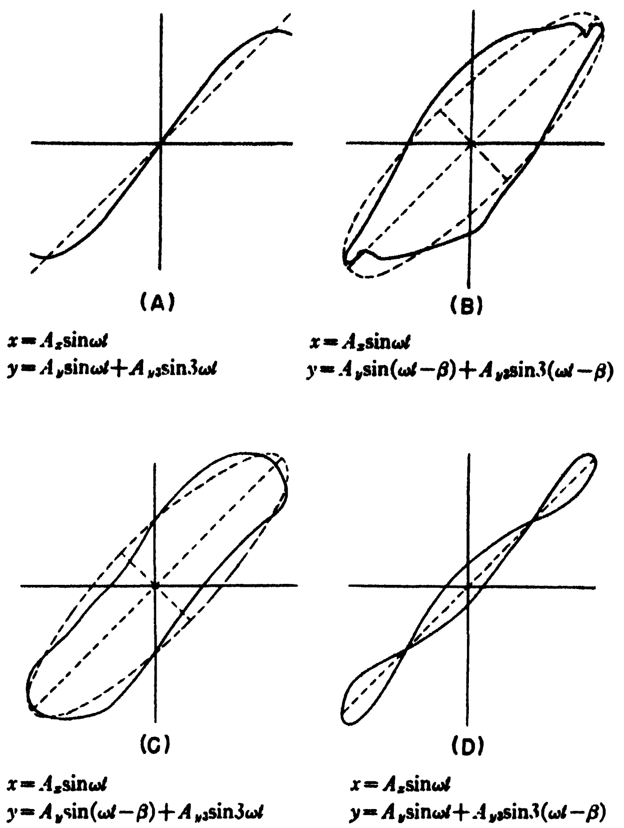
Percentage of second harmonic =

$$\frac{m_1 + m_2}{4\,A_y \cos\gamma} \times 100 \qquad (12\text{-}3)$$

where the symbols are as indicated in Fig. 12-31A. As is seen from the figure, A_y is obtained by measuring the vertical height of the pattern, while m_1 and m_2 are the maximum widths of the upper and lower lobes, respectively. When both signals are in phase, including the harmonic of the y signal, m_1 will be equal to m_2.

For the pattern at part (B) of Fig. 12-31, both the fundamental and second harmonic of the y input signal are in phase with each other, but each is out of phase with the x input signal by the angle β. For the pattern of part (C), only the second harmonic of the y-input signal is in phase with the x-input signal, whereas the fundamental of the y-input signal is out of phase with the x signal by the angle β. In part (D) the pattern results when the fundamental of the y-input signal is in phase, but where its second-harmonic component is out of phase by the angle β, with the x-input signal.

The four patterns of Fig. 12-32 result when the input signal to the vertical-deflection plates contains only a fundamental plus a third harmonic. The phase between the fundamental and third harmonic of the y or vertical input signal and the x or horizontal input signal for parts (A) to (D) of Fig. 12-32 is the same as that for the respective patterns of Fig. 12-31. This can easily be verified by examination of the x and y equations for each pattern.



(A) $x = A_x \sin\omega t$, $y = A_y \sin\omega t + A_{y3} \sin 3\omega t$

(B) $x = A_x \sin\omega t$, $y = A_y \sin(\omega t - \beta) + A_{y3} \sin 3(\omega t - \beta)$

(C) $x = A_x \sin\omega t$, $y = A_y \sin(\omega t - \beta) + A_{y3} \sin 3\omega t$

(D) $x = A_x \sin\omega t$, $y = A_y \sin\omega t + A_{y3} \sin 3(\omega t - \beta)$

Courtesy Rev. Sci Instr.

Fig. 12-32.—Similar to Fig. 12-31 except that only a third harmonic is contained in the complex vertical-deflection signal.

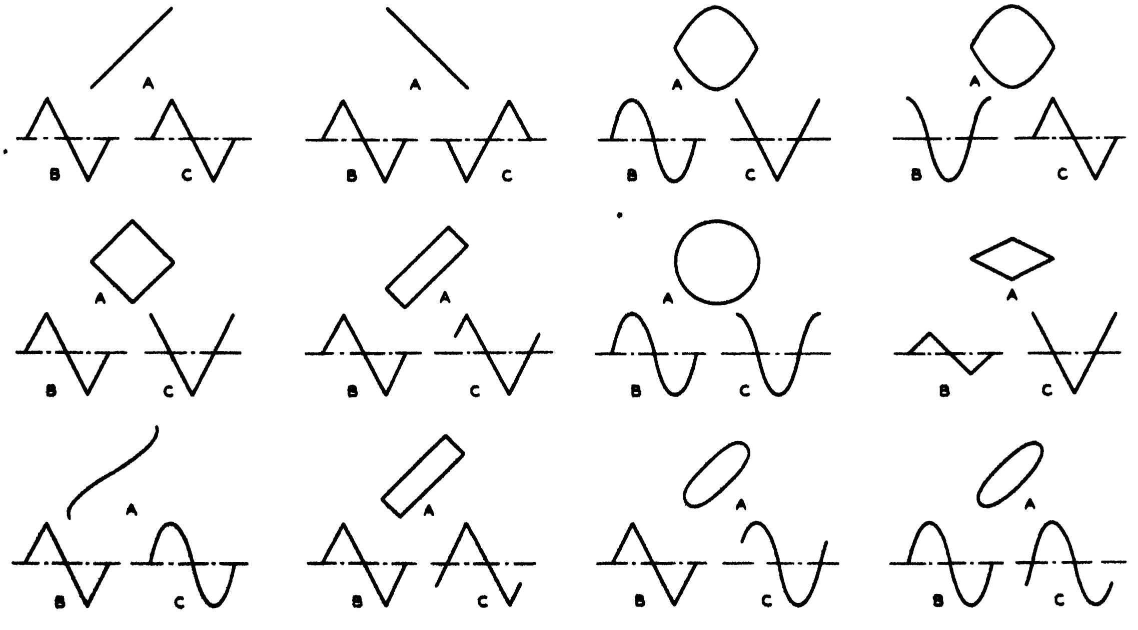


Fig. 12-33.—In each group *A* represents the resultant image, *B* the vertical-deflection voltage, and C the horizontal-deflection voltage. The amplitude and also the frequency of the *B* and *C* signals are the same in each grouping.

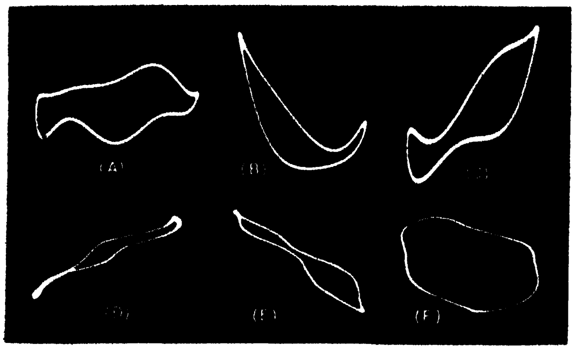


Fig. 12-34.—Various patterns resulting from 1:1 frequency ratios where either one or both of the input signals are distorted.

A general idea of the effect of waveshape and phase relation upon the pattern developed is indicated by the various drawings of Fig. 12-33. In each of these, *A* is the resultant image, *B* the vertical-deflection voltage, and *C* the horizontal-deflection voltage. The frequency, as well as the amplitude, of each *B* and *C* input signal is the same. From these patterns, one can gather an idea of the resultant patterns, under the conditions cited, when the input signals have waveshapes which lie between limits represented by sine waves at one extreme and rectilinear waveforms at the other limit. Each resultant pattern can easily be constructed by a graphical analysis similar to that used for Fig. 12-20.

The patterns of Fig. 12-34 represent conditions where either one or both of the input signals to the deflection plates are distorted, but where the frequencies of the input signals are equal. Although these 1:1 frequency-ratio patterns are all shown with some degree of opening, other 1:1 patterns may take the form of a single crooked line to various distorted elliptical shapes. For the patterns at (A), (B), and (C) of Fig. 12-34, the input to the horizontal plates is a sine wave, while that to the vertical plates is a distorted wave. For (A), the frequency of both input signals is 1,000 cps while for (B) and (C) both frequencies are 1,500 cps. The patterns at (D), (E), and (F) were taken where both input signals were distorted.

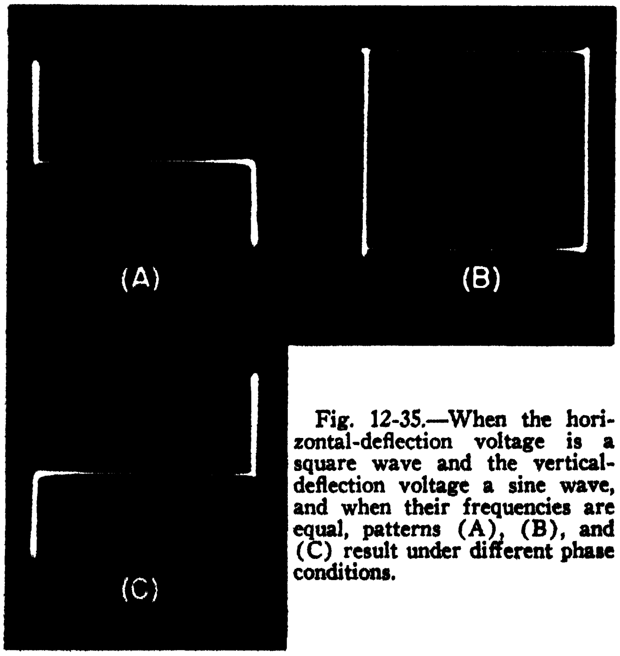


Fig. 12-35.—When the horizontal-deflection voltage is a square wave and the vertical-deflection voltage a sine wave, and when their frequencies are equal, patterns (A), (B), and (C) result under different phase conditions.

The three 1:1 ratio patterns of Fig. 12-35 are produced when the input signal to the horizontal-deflection plates is a (1,500 cps) square wave and the input to the vertical-deflection plates is a (1,500 cps) sine wave. The pattern at (A) indicates a 0° phase difference, that at (B) is said to represent a 90° or 270° phase difference, and the pattern at (C) is a 180° phase difference. A square pattern effect is obtained for a 90° or 270° phase difference, as seen in part (B), when the amplitudes of both input signals are equal.

Determining the Motion of the Spot

In previous sections, we have seen how important it is to know in which direction the spot is traveling in order to determine what phase difference is represented by the pattern in question. In this section, we will study a very simple method that will show us the direction of motion of the spot at all times. This method involves the use of a special waveform applied to the grid of the cathode-ray tube, which will modulate the intensity of the electron beam. A number of different circuits[4] have been developed by the authors in the laboratory of the publisher and each has been found to produce satisfying results. In this section, we will study this special waveform from the *ideal* viewpoint and then show oscillograms indicating the direction of spot travel as produced by one of these circuits.

If any signal is fed to the grid circuit of a cathode-ray tube, it will vary the bias already present on the grid and hence modulate the intensity of the electron beam. A form of intensity modulation was already treated in Chapter 10 under the discussion of blanking. If we were to inject a square wave into the grid of a cathode-ray tube, it would cause the intensity of the electron beam to increase on the positive half-cycles and decrease on the negative half-cycles. If the amplitude of the square wave is strong enough, it may, during its negative half-cycles, drive the grid of the cathode-ray tube beyond its cutoff region, and thus cause blanking during these periods. The square wave shown in Fig. 12-36A has its negative half-cycles greater than the cutoff bias of the cathode-ray tube and thus will cause blanking of the electron beam during these periods.

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[4] All circuits and processes relative to the use of such a system have patent applications pending.

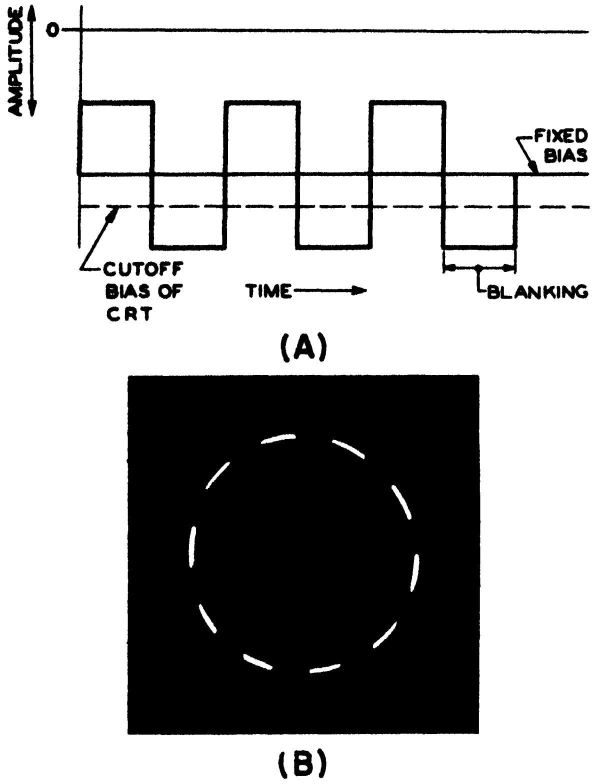

Fig. 12-36.—When the square wave at (A) is fed to the grid of a cathode-ray tube which has a circular trace on the screen it will intensity-modulate the trace, as indicated at (B).

Assume two sine waves 90° (or 270°) out of phase and equal in amplitude and frequency (500 cps each). If one is applied to the vertical plates and the other to the horizontal plates, the resulting pattern will be a circle. If a square wave similar to Fig. 12-36A and at a frequency of 5,000 cps is applied to the grid of the cathode-ray tube to intensity-modulate[5] the circular image, the resultant pattern will appear as shown in Fig. 12-36B. From this photograph we see that the periods of blanking and those of the traces are approximately equal in length and that there are ten periods each of blanking and trace. The direction of spot travel cannot be determined from the present pattern, however.

The method of determining the direction of spot travel is predicated upon the addition of a pulse to the lagging edge of positive half-cycles of a square wave. This is indicated by the ideal waveshape of Fig. 12-37. The repetition frequency of the pulse should be the same as that of the square wave. Pulse additions are only indicated on the positive half-cycles of the square wave, but actually any type of addition may be made along the negative half-cycles of the square wave. As long as the waveshape during the negative half-cycles is always more negative than the cutoff bias, the exact form of the waveshape of these negative half-cycles is not important. This is so because these portions of the signal will always produce blanking.

If a signal that has positive half-cycles, similar to that of Fig. 12-37, is applied to the grid of a cathode-ray tube, the resultant pattern will contain alternate blanking intervals and traces, as in the pattern of Fig. 12-36B, but from the shape of the visible parts of the pattern, the direction of spot travel is noticed readily. Since variation of the grid bias controls the intensity of the electron beam, the less negative the grid bias, the greater is the intensity, and the more negative the bias, the less is the intensity. (This was indicated in Chapter 10 under the discussion of the *intensity control.*) Consequently, for any one setting of the intensity control, those parts of the positive half-cycles of the signal shown in Fig. 12-37 from time *A* to *B* will produce a trace upon the screen at one constant intensity, and from time *B* to *C* the trace will be at a greater intensity. For the negative half-cycles of the signal of Fig. 12-37 the grid is driven beyond cutoff and blanking occurs.

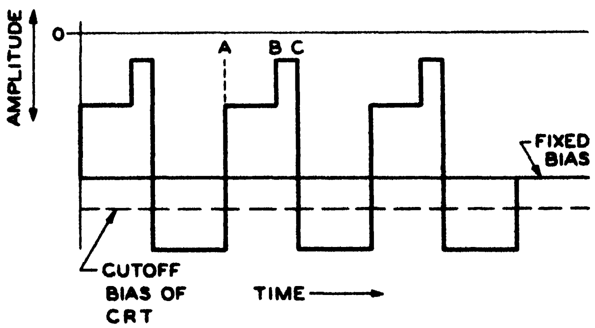

Fig. 12-37.—A positive pulse of the same frequency as the square wave is added to the lagging edge of each cycle of square wave.

We know that for high intensities of the beam, the trace or pattern becomes thicker during such times. This means that during the visible portions of the pattern one end of each individual trace (between blanking intervals) is thicker than the rest of the trace. If the pulse width is small compared to the width of the half-cycle of the square wave, as shown in Fig. 12-37, the visible portions of the pattern will each appear as a trace with a *dot* at one end. *If the dot is considered as an arrow head and the remainder of the trace as the tail of the arrow, then we have a ready means of telling the direction of spot travel.*

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[5] Intensity modulation is also used as a means of frequency measurement and is discussed in greater detail later on in this chapter.

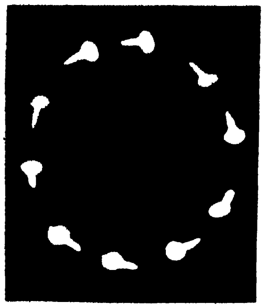

Fig. 12-38.—When a 5,000-cps signal similar in shape to that of Fig. 12-37 is fed to the grid of the cathode-ray tube, the circular trace which is formed by 500-cps deflection signal will be intensity-modulated as shown.

Let us now see how the circular pattern changes when a signal similar to that of Fig. 12-37 is applied to the grid of the cathode-ray tube. The resultant pattern in question is illustrated in Fig. 12-38. The two input signals to the deflection plates have the same frequency as before, namely 500 cps, and the frequency of the special wave input is 5,000 cps so that there will be ten blanking intervals along with ten visible traces of the pattern (the same number as in Fig. 12-36B). From the resultant pattern of Fig. 12-38 we see that the *dot* or head of each visible trace is oriented in the clockwise direction. Since the spot travels in a clockwise direction, the phase difference represented by the pattern is undoubtedly 90°. Since we saw previously that for angles of *lead* less than 180° of the vertical- with respect to the horizontal-deflection voltage, the pattern will be traced in a clockwise direction.

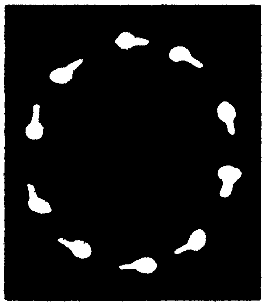

Fig. 12-39.—The dots on each visible trace in this 1:1 frequency ratio pattern is oriented in the counterclockwise direction indicating a 270° phase difference.

For a phase difference of 270°, the spot would travel in a counterclockwise direction. This is indicated by the pattern of Fig. 12-39 which was produced by the same conditions as Fig. 12-38, except that one of the two input signals was advanced 180° with respect to the other. If the input signal to the vertical-deflection plates *leads* that to the horizontal-deflection plates by 90°, the spot will trace out the circle in a clockwise direction. However, if the vertical-deflection signal *lags* the horizontal-deflection voltage by 90°, the spot will trace out the circle in a counterclockwise direction. When we say that a circular pattern represents a phase difference of 270°, for the counterclockwise case, we mean that the vertical-deflection voltage is *leading* the horizontal-deflection voltage by 270°. This is the same as saying that the horizontal input signal leads the vertical by 90°, and occasionally may be expressed more conveniently in this way.

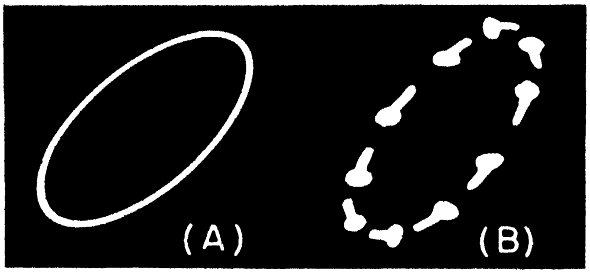

Fig 12-40.—The ellipse at (A) can represent either a 45° or 315° phase difference but the intensity-modulated pattern at (B) indicates the difference to be 315°.

Let us study some more phase patterns which are intensity-modulated with the special signal so as to determine the direction of spot travel. In Fig. 12-40 appear two photos. That at (A) is an ellipse which can approximately represent either a 45° or 315° phase relationship. The pattern at (B) shows the direction of spot travel for the pattern at (A) to be counterclockwise, which tells us that the phase difference is 315°. The ellipse of Fig. 12-41A approximately indicates either a 135° or 225° phase difference and the pattern at (B) definitely establishes this phase difference to be 135°, because the spot is traveling in a clockwise direction.

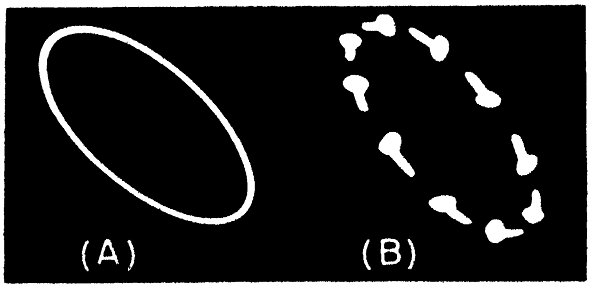

Fig. 12-41.—The ellipse at (A) indicates either a 135° or 225° phase difference but the intensity-modulated pattern at (B) establishes the difference as 135°.

Thus far we have seen how the pattern looks under this special condition of intensity modulation, when the spot travels in one direction only and *does not* trace back on itself. We have previously shown, by graphical means, that for straight-line phase-difference patterns of 0° and 180° (see Figs. 12-3 and 12-10, respectively) the spot actually traces over the same pattern twice, but in opposite directions. On this basis of spot travel, the straight line, when intensity-modulated with the special waveform, should have dots at *both ends* of each visible trace. This is brought out by the photographs of Fig. 12-42. Part (A) of this figure represents a 0° phase-difference pattern and part (B) a 180° difference. Note that, in both patterns, each trace has a head on both ends, indicating that the spot traces each straight line twice, but in opposite directions. For this particular case, the frequency of the grid-modulating signal was eight times that of the input signals.

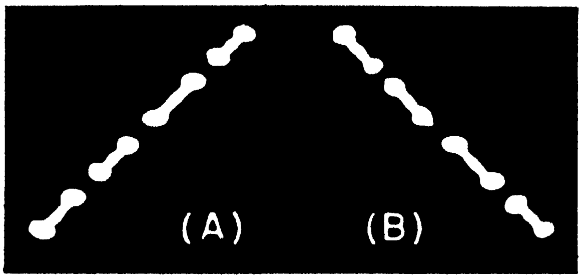

Fig. 12-42.—The dots at both ends of each trace of these straight-line images indicates that the lines are traced out twice.

With the ability to determine the direction of motion of the spot, the teaching of cathode-ray-tube principles, relative to beam deflection and patterns portrayed, becomes a very simple matter. By employing special waveforms as have been discussed here for intensity modulation, the oscilloscope becomes more useful for educational and demonstration purposes. The circuits that were designed to produce this special waveform can be incorporated as an integral part of each oscilloscope or can be used as an auxiliary piece of equipment.

Stationary Patterns

Since phase-angle tests are made with a single source of voltage, the frequency of the voltage applied to both sets of deflection plates must be constant so that the phase difference is constant, at whatever value it may be. Consequently, the pattern will remain stationary. Whether or not the patterns will remain stationary for any one condition, or will drift from shape to shape, that is, drift from a single line to an ellipse and to a circle and reverse, depends upon the frequency stability of the input signals. If the frequency of one of the deflection voltages is continuously changing, no matter how slight this drift, the pattern will not remain stationary. It is going to change its shape continually, because the phase relation is continually changing. Normally, two signals which bear a certain phase relation to each other remain in that phase relation as long as their relative frequencies and other controlling factors remain constant. The movement of a pattern resulting from the application of two frequencies, which are substantially identical, takes place at a rate equal to their difference frequency. In other words, the rate of change of a pattern through the complete 360° cycle is an indication of the frequency difference between the two frequencies.

Frequency Comparisons

From what has been said so far, it is evident that these phase patterns are also suitable for frequency comparison. In other words, a stationary pattern which is a single line, circle, or ellipse, shows that both input frequencies are identical, and we say that they have a 1 :1 frequency ratio. If one of the voltages is of known frequency and the other unknown, this unknown frequency may be calibrated against the known by mutual adjustment to produce one of the patterns mentioned. In this connection, the cathode-ray tube affords an excellent and extremely accurate means of comparing audio frequencies which normally require very elaborate and expensive standards. Frequency comparison can also be made when the frequency ratio is other than 1 :1. Regardless of the frequency relationship, the resulting pattern is always known as a Lissajous figure. For the pattern to be stationary, the frequencies to be compared must exhibit a simple integral (whole number) relationship with respect to each other.

Lissajous patterns can be developed under various conditions and they will be of diverse shapes. They may be relatively complex or simple, depending upon the frequency ratio between the two varying voltages, their amplitude and phase relations, waveshape, and respective modulation. As a general rule, when used for the purposes of frequency comparison, the known signal is secured from a sine-wave oscillator, so that a certain expected type of pattern is usually developed.

We have shown both by geometrical analysis and practical oscillograms what patterns result from the application of two sine waves of equal frequency. Let us now see, by similar methods, what the resulting pattern will be when the input signals have integral frequency ratios other than the simple 1 :1 ratio. In the graphical illustrations which follow, the input signals will again be sine waves. The signals in question can either be fed through the vertical and horizontal amplifiers of the oscilloscope, through external amplifiers, or applied directly to the deflection plates of the cathode-ray tube. In the latter case, the signals have to be strong enough to produce a suitable deflection of the beam. Whatever system is actually used, the relative phase difference and amplitude relationship between the two input signals will be assumed to exist *directly* at the deflection plates in the following discussion.

A 2:1 Lissajous Pattern — 0° and 180° Phase Difference

The first illustration will be for a frequency relationship of 2:1. Let us discuss the case where the input signal to the vertical-deflection plates is twice the horizontal input frequency. For the sake of illustration, both signals have equal amplitudes and are in phase. The geometrical drawing of Fig. 12-43 is constructed similarly to that shown for the 1:1 ratios. The exact frequencies are of no importance for this discussion, as long as they fall within the frequency response of the oscilloscope amplifiers.

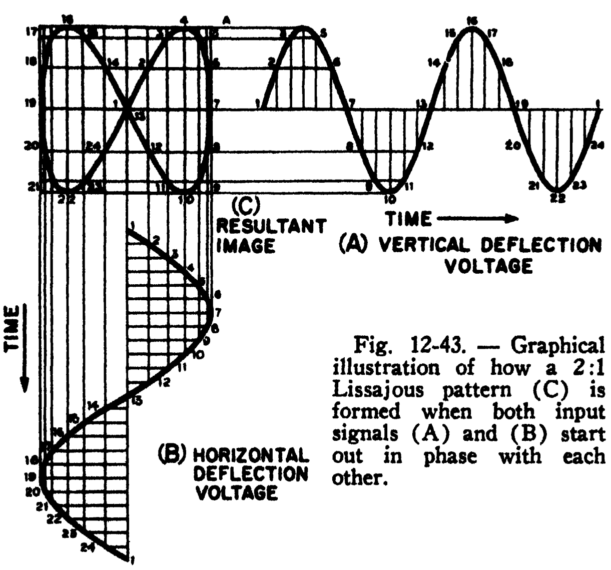
(C) RESULTANT IMAGE
TIME →
(A) VERTICAL DEFLECTION VOLTAGE
TIME
(B) HORIZONTAL DEFLECTION VOLTAGE

Fig. 12-43. — Graphical illustration of how a 2:1 Lissajous pattern (C) is formed when both input signals (A) and (B) start out in phase with each other.

The sine wave at (A) is the vertical-deflection voltage, that at (B) the horizontal-deflection voltage, and that at (C) the resultant image. The numbers on the resultant pattern correspond to those on each sine wave. The two-cycle sine wave at (A) is divided into 24 equal time segments of 30° each, and in order to divide the sine wave at (B) into the same number of segments, each segment for this single cycle is 15° apart. Thus we see that every time the vertical input signal advances by 30°, the horizontal input signal advances by exactly half that amount, namely by 15°.

The resultant pattern is, as you will see, in the shape of a symmetrical figure eight horizontally situated. We want to call to your attention once again that the pattern is made by a single spot which moves through these various positions. The fact that the image appears as a stationary pattern is due to the repeated movement of the spot through the same positions. The pattern is completely symmetrical about both the horizontal and vertical axes. Tangents drawn to each side would produce a square, because the amplitudes of the deflection voltages are equal. If the vertical-deflection voltage were greater than the horizontal-deflection voltage, the height of the resultant pattern would increase and tangents drawn to each side would produce a rectangle with a height greater than its width. Conversely, if the horizontal voltage were greater than the vertical-deflection voltage, the pattern would widen and become flatter.

From the drawing of Fig. 12-43, both input signals are in phase and the spot traces out the resultant pattern in the direction of increasing numbers. Thus the spot moves from the origin into quadrant I and continues around the pattern as shown. If the phase of the input signal to the horizontal-deflection plates were advanced 180°, exactly the same pattern would result and the direction of spot travel would also be the same.

Here we have an unusual case: for two initial phase differences, one 0° and the other 180°, the spot travels in the same direction and traces out the same shape pattern. There are still further interesting details concerning the phase relationship between the two input signals. In Fig. 12-43, signals (A) and (B) are in phase and the voltages of both increase in the *positive* direction from their baselines. However, if the two input signals are applied at an instant where both voltages start increasing in the *negative* direction, the same shape pattern will result, but the spot will be traveling in the *opposite* direction in tracing out this pattern. The spot would start traveling from the origin into quadrant III, instead of quadrant I.

If only the vertical-deflection voltage of Fig. 12-43 were changed, so that it would increase in a negative direction below the baseline, while the horizontal voltage were increasing in the positive direction, once again the same pattern would result. The direction of spot travel for this case would be the same as for the case where the signals both start to increase in the negative direction.

The photograph of Fig. 12-44A illustrates a frequency ratio of 2:1 and is similar to that constructed by the graphical analysis of Fig. 12-43. This photograph was taken, however, with the horizontal-deflection voltage being less than the vertical-deflection voltage.

Refer to Fig. 12-44B. This pattern is the same as part (A), except that the special grid waveform has been applied to determine the direction of spot travel.

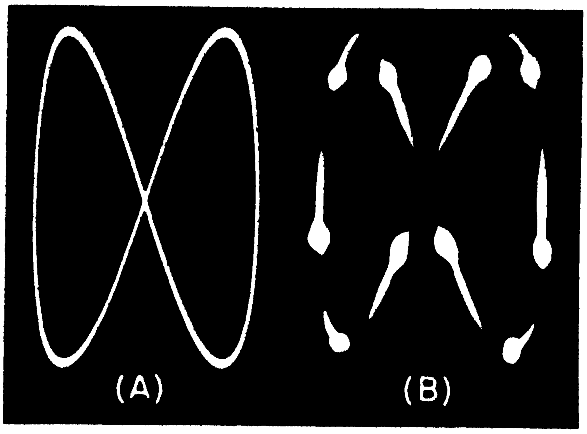

Fig. 12-44.—The Lissajous pattern at (A) is for a 2:1 frequency ratio and may indicate an initial phase condition of either 0° or 180°; the pattern at (B) indicates the phase condition to be 180°.

From this photograph, the direction of spot travel is as shown in Fig. 12-43, or where the initial phase of the horizontal input voltage is shifted by 180°.

If the phase relation between the two input signals is other than the cases discussed, the pattern for a 2:1 frequency ratio would be different from that indicated in Fig. 12-44A. Let us graphically analyze some of the resulting patterns for varying phase conditions.

Other 2:1 Patterns

Fig. 12-45 is a series of graphical constructions depicting various Lissajous patterns for a frequency relationship of 2:1 under different phase conditions. The signals at (A) represent the vertical-deflection voltages, those at (B) the horizontal-deflection voltages, and those at (C) the resultant image. Resultant pattern *C1* is the same as that for Fig. 12-43 and is included with the others for comparison purposes. The degree markings next to each resultant image indicate the possible initial phase relations of the vertical-deflection voltage to produce the pattern shown; the horizontal input is kept the same throughout. The construction was made for input signals of equal amplitude.

As the phase relation changes from 0° to 90°, the loops of the pattern gradually close (i.e., the area enclosed by each loop becomes less). At 90°, the loops become completely closed as seen from pattern *C3* in Fig. 12-45. The spot actually traces through this pattern twice. From 90° to 180°, the pattern opens up again. At 120°, pattern *C2* is repeated again, and at 180°, pattern *C1* is repeated. It is seen that from 0° to 180° the resultant patterns can be thought of as being *cup-shaped downward.*

From 180° to 270°, the loops gradually become closed again, but the patterns reverse themselves. At 210° phase relation, the pattern is as shown at *C4* in Fig. 12-45. At 270°, pattern *C5*, the loops are completely closed; again, the spot traces this pattern twice.

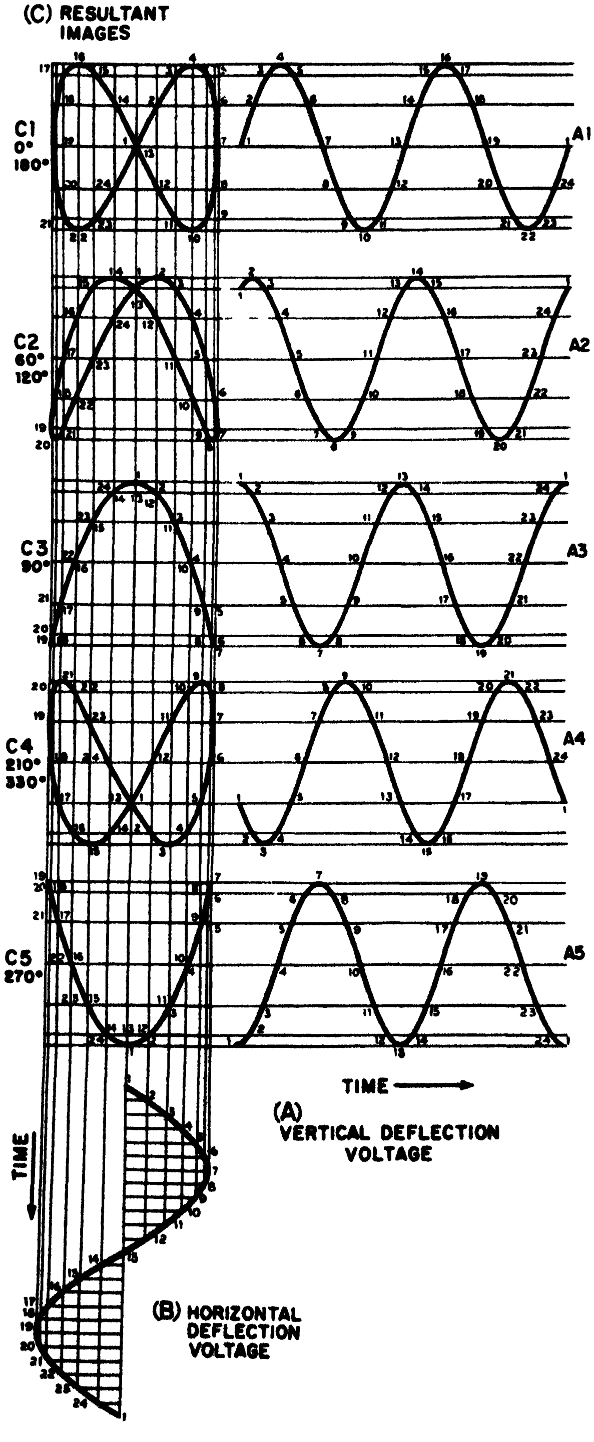

Fig. 12-45.—Various 2:1 frequency ratio patterns are shown at (C) for different phase conditions between the two input signals (A) and (B).

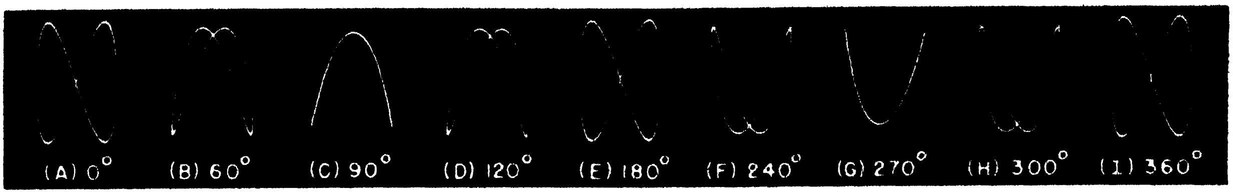

Fig. 12-46.—A series of 2·1 Lissajous patterns where initial phase angle of the vertical-deflection voltage is as shown and where the higher-frequency input is to the vertical-deflection plates.

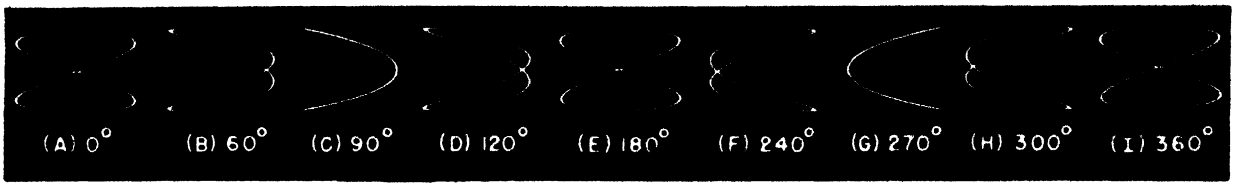

Fig. 12-47—Keeping all conditions producing the patterns of Fig. 12-46 the same but only reversing the input signals to the deflection plates produces the series of 2.1 Lissajous patterns shown here.

From 270° to 360° the loops open up again. At 330° the pattern becomes the same as that at 210° (see resultant image *C4*), and in approaching 360° the pattern approaches that of *C1* again. After 180°, the resultant patterns may be said to be *cup-shaped upward.*

A number of oscillographic waveforms of 2:1 Lissajous patterns were photographed and are shown in Fig. 12-46. The higher-frequency input is to the vertical-deflection plates. The patterns from (A) to (I) indicate increasing initial phase angles for the vertical-deflection voltage, while keeping the horizontal input unchanged. Note that the patterns at (A), (E), and (I) are of the same shape. The patterns at (B) and (D) and those at (F) and (H) are also the same. The patterns at (C) and (G) each occur at only one phase difference, namely 90° and 270°, respectively.

If the signal inputs to the cathode-ray tube were reversed, that is, if the higher-frequency signal were to cause the horizontal deflection and the lower-frequency signal the vertical deflection, the resulting patterns would be similar to those of Fig. 12-46, but they would be turned through 90° in the clockwise direction. The nine photos of Fig. 12-47 indicate various patterns when the horizontal input signal is twice the frequency of the vertical input signal. The initial phase of the horizontal voltage for a constant vertical input voltage is marked underneath each photo. The phase relations for the photos from (A) to (I) of Fig. 12-47 were chosen to be the same as those of parts (A) to (I) of Fig. 12-46, respectively.

Before we discuss the method of determining the frequency relationship from the patterns, let us study, by similar geometrical construction, the type of patterns formed when the frequency ratio is 3:1.

3:1 Lissajous Patterns

In the analysis to follow, the frequency of the signal input to the vertical-deflection plates is three times as great as the signal input to the horizontal-deflection plates. The construction is shown in Fig. 12-48. Part (A) is the vertical-deflection voltage, part (B) the horizontal-deflection voltage, and part (C) the resultant image. The initial phase of signal (A) is 90° because it starts at a maximum positive voltage, whereas signal (B) has an initial phase angle of 180°, because it starts at a zero voltage and increases in the negative direction.

This departs from our previous practice of starting both the horizontal- and vertical-deflection voltages in the positive direction from the baseline, but it will serve to bring out the fact that the same resultant patterns will be produced, no matter what the initial phase angles are. If (B) were increasing in the positive direction, the same pattern would result, but it would be traced out in the opposite direction, as is easily seen from the figure.

The pattern is constructed in the same manner as the previous examples. Both vertical- and horizontal-deflection voltages are divided into the same equal time intervals. To develop the pattern, the spot is positioned as a result of the amplitude displacement of both waves. These spot positions are indicated by the same numbers for each wave. For the case illustrated, the amplitudes of both signals are equal. Thus tangents drawn around the resulting pattern would form a square. If the horizontal-deflection voltage were greater than the vertical voltage, the height of the pattern would be larger than the width. Conversely, if the vertical-

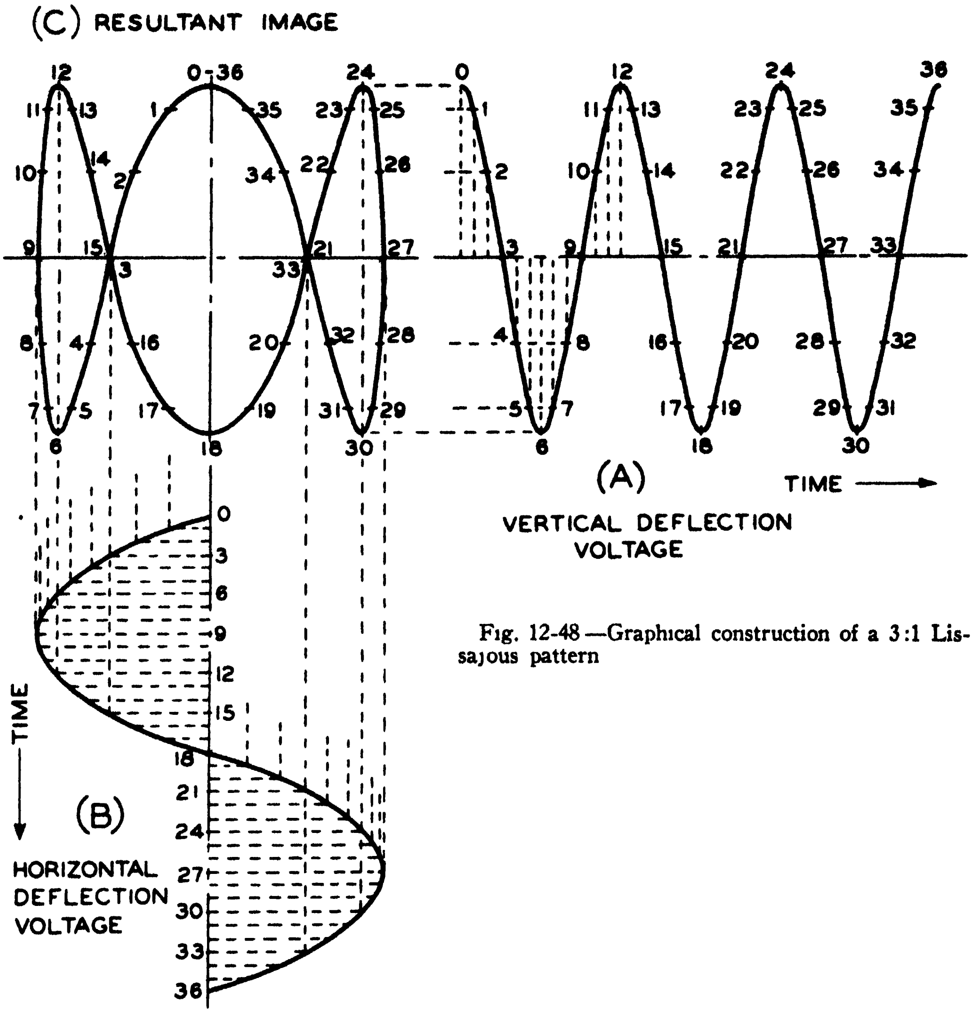

Fig. 12-48—Graphical construction of a 3:1 Lissajous pattern

deflection voltage were greater, the width of the resulting pattern would be larger than the height. Note that the pattern contains three loops.

For comparison purposes, the input signals and the Lissajous pattern of Fig. 12-48 were reproduced on the oscilloscope and then photographed. The resulting oscillograms were arranged to conform to that of Fig. 12-48 and are illustrated in Fig. 12-49. Parts (A), (B), and (C) of Figs 12-48 and 12-49 are the same. If the starting phase of either input signal (A) or (B) of both figures were shifted by 180°, the resultant pattern would still be the same.

It should be remembered that these phase angles mentioned for Lissajous patterns greater than a 1:1 ratio are only relative to the initial starting conditions of the two input signals as shown graphically. Actually, to speak of a "phase difference" between two signals of different frequencies is not meaningful, since the relative phase will not be the same at every instant of

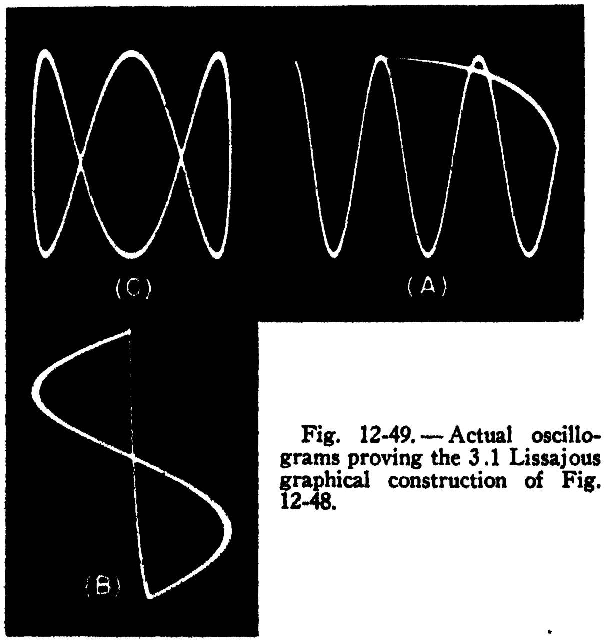

Fig. 12-49.—Actual oscillograms proving the 3.1 Lissajous graphical construction of Fig. 12-48.

time. To repeat, the stipulated phase relations in Fig. 12-48 mean that signal (A) starts out at a maximum positive amplitude at the time signal (B) starts out at zero voltage. Whether signal (B) starts to increase in the positive or negative direction makes no difference in the final shape of the pattern; all it means is that the spot travels in opposite directions in tracing out the pattern.

In Fig. 12-50 appears another graphical analysis of a 3:1 frequency ratio, but in this case, the higher-frequency vertical input signal (A) and also the horizontal input (B) both start out at zero voltage; however, signal (B) has an initial phase angle of 180°. If both input signals were shifted in initial phase by 180°, the same pattern would result. If only one of the input signals were shifted in its starting phase by 180°, a different pattern would result because, although both signals would start out at zero voltage, they would increase in the same direction. In this latter instance, the pattern will be reversed; that is, instead of the peaks of the pattern in Fig. 12-50C being in quadrants II and IV, they will be in quadrants I and III.

The transition from the pattern of Fig. 12-48 to that of Fig. 12-50 occurs in the time required for the vertical-deflection signal to shift in phase by 90° in the lagging direction or 270° in the leading direction. The two Lissajous patterns illustrated in Fig. 12-48 and Fig. 12-50 represent the extreme changes in shape possible for 3:1 frequency ratios. As the starting phase of one input signal varies (in our illustrations it is the vertical-deflection signal), the loops of the resultant image of Fig. 12-48C start to close until at Fig. 12-50C they are completely closed.

To comprehend properly such Lissajous figures, it is necessary to imagine the pattern of Fig. 12-48C traveling around a transparent glass cylinder, through which the front and rear patterns are both visible. The view obtained is the elevation or side view. At certain time intervals of the moving pattern, some parts of the pattern would be directly behind each other, and, there-

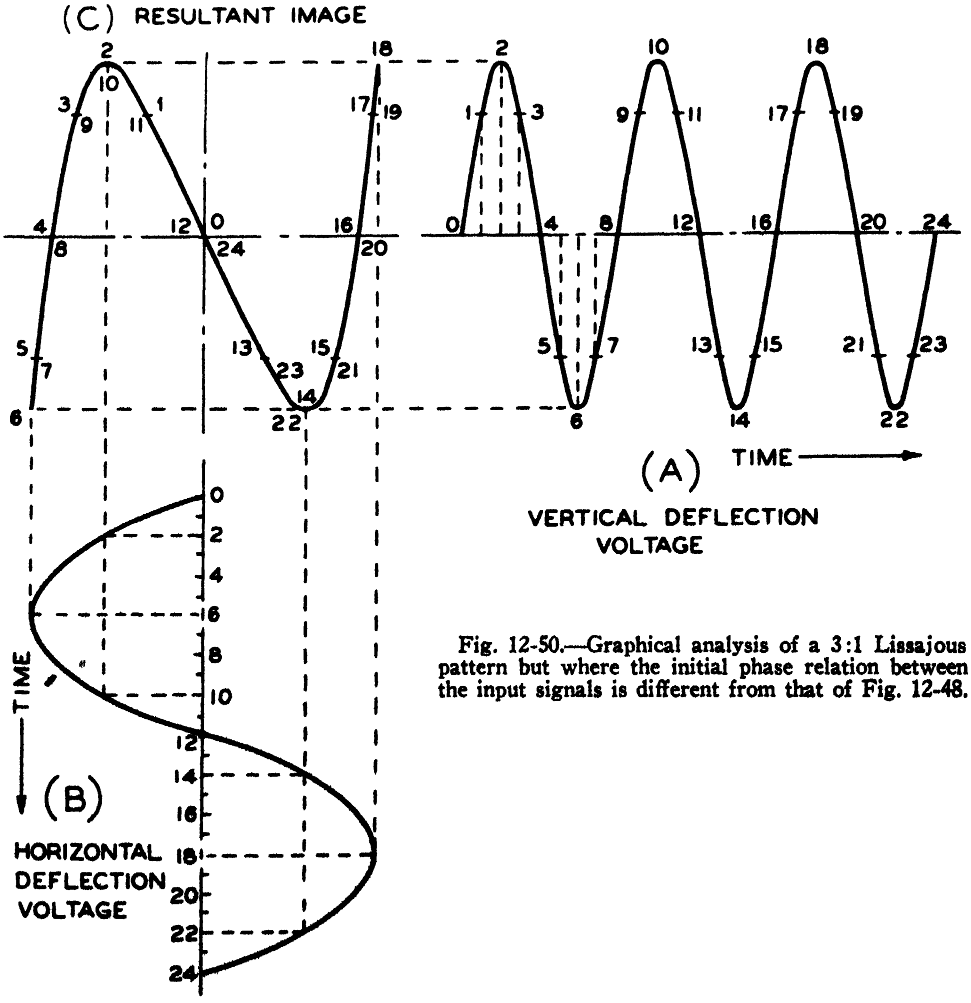

Fig. 12-50.—Graphical analysis of a 3:1 Lissajous pattern but where the initial phase relation between the input signals is different from that of Fig. 12-48.

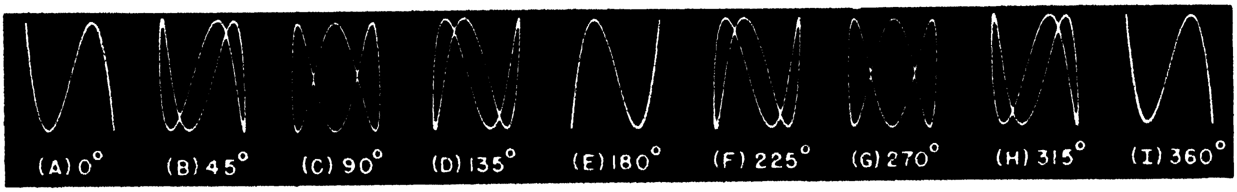

Fig. 12-51.—A series of 3:1 Lissajous patterns for different initial phase conditions of the vertical-deflection voltage which is the higher-frequency input signal.

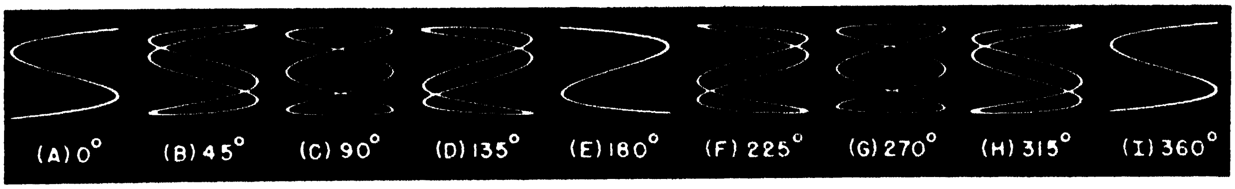

Fig. 12-52.—When the horizontal-deflection voltage is the higher-frequency signal, the 3:1 Lissajous patterns of Fig. 12-51 change to those shown here.

fore, would not be visible. This is also true for the 2:1 Lissajous patterns previously discussed, as well as for all other ratios.

In Fig. 12-51 appear a series of Lissajous patterns, each indicating a 3:1 frequency ratio with the vertical-deflection signal being the higher frequency. These photos indicate, in progressive fashion, how the pattern changes when the vertical-deflection voltage changes its initial phase from 0° to 360° with respect to a starting position of zero, and the amplitude increasing in the positive direction from the baseline. The horizontal-deflection voltage is assumed to be in the same starting position (0° initial phase), and is not changed, while the phase of the vertical voltage is varied through 360°. The phase change is in approximate steps of 45° as indicated under each photo. Note that patterns (A) and (I) are the same, (B) and (H) are the same, (C) and (G) are the same, (D) and (F) are the same, and that only (E) has no duplicate. Since 360° coincides with 0°, pattern (A) is really also unique.

If the higher-frequency input signal were impressed across the horizontal-deflection plates, the resulting 3:1 ratio patterns of Fig. 12-51 would be repeated, but they would again be rotated clockwise through 90°. This is indicated by the Lissajous patterns of Fig. 12-52. Parts (A) to (I) of this drawing are counterparts of the same letters and initial phase angles of Fig. 12-51.

3:2 Lissajous Patterns

Thus far we have studied Lissajous patterns of 2:1 and 3:1 frequency ratios. The same patterns will also result if the ratios are 4:2 or 6:2 respectively, because each of these latter ratios can be simplified to the former. In such ratios as these, one frequency is an integral multiple of the other. Let us now study some Lissajous patterns where one frequency is *not* an integral multiple of the other. Thus if the higher frequency is expressed as the numerator and the lower frequency as the denominator, the resulting ratio will always be fractional and cannot be expressed as a whole number.

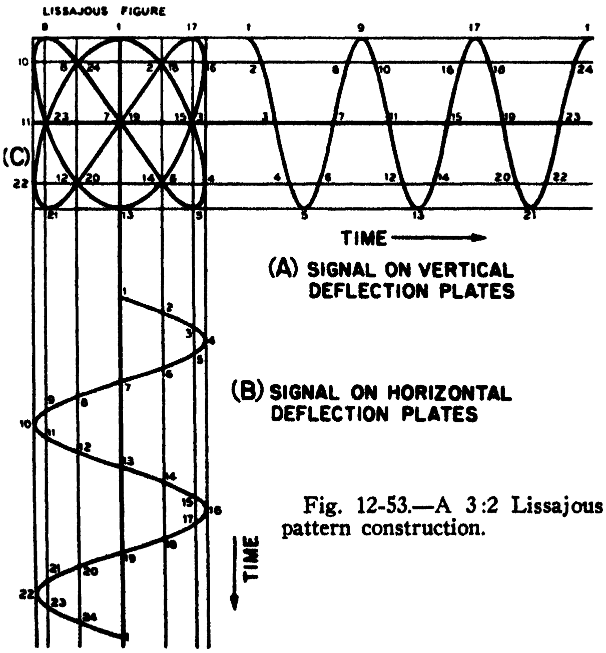
LISSAJOUS FIGURE
(C)
TIME
(A) SIGNAL ON VERTICAL DEFLECTION PLATES
(B) SIGNAL ON HORIZONTAL DEFLECTION PLATES
TIME

Fig. 12-53.—A 3:2 Lissajous pattern construction.

In Fig. 12-53 appears a graphical construction of a 3:2 Lissajous pattern. In the case under illustration, (A) is the vertical-deflection voltage and is the higher-frequency signal, (B) is the horizontal-deflection voltage, and (C) is the resultant pattern. The relative phase position is such that the signal at (A) is *initially* leading that at (B) by 90°. The construction is along the same lines as those of the previous drawings. The time axis for each input signal is divided into 24 equal segments and indicated by the same numbers.

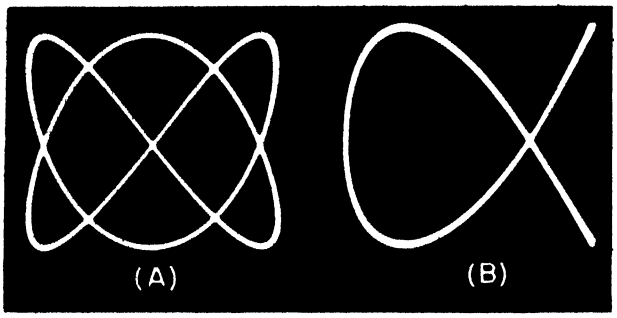

Fig. 12-54.—Pattern (A) is an oscillogram indicating the 3:2 ratio but for different initial phase conditions between the two input signals.

The resultant pattern is completely "opened." By changing the initial phase of either signal by an appropriate amount, the spot can be made to trace back on itself so that the resulting pattern will actually consist of two coincident lines (as was the case for Figs. 12-46C and G, and 12-51A and E.) In Fig. 12-54 appear two oscillograms of patterns depicting 3:2 frequency ratios. That at (A) is produced by the input signals as shown in Fig. 12-53. That at (B) results when the horizontal input signal of Fig. 12-53 is shifted in phase by 90° (in a leading direction). This latter photo consists of two coincident lines, because the spot traces the pattern twice in opposite directions.

Classification of Lissajous Patterns

From the analysis of Lissajous patterns (based on sine-wave input) thus far, we can classify the patterns and make some generalizations. We do know that it is the initial phase of the two input signals that determines the final shape of the pattern. We also know that in shifting the phase of one signal, keeping the other stationary, we can change the pattern consisting of *completed loops,* such as those of Figs. 12-46A and B, 12-51B and C, and 12-54A, to those in which the pattern is actually a *double image* appearing as a single line. The latter types of patterns, which consist of two coincident lines, are represented by the photos of Figs. 12-46C and G, 12-51A and E, and 12-54B. The *double-image* patterns are often classified as *uncompleted loop or closed* patterns.

In those double-image or uncompleted loop patterns where the higher frequency is an integral multiple of the lower frequency (i.e., 2:1, 3:1, 4:1, 5:1, etc.), there are no intersecting lines and hence no completed loops (See Figs. 12-46C and G and 12-51 A and E). In those double-image patterns, where the frequency ratio in its simplest form is represented as a fractional equivalent (i.e., 3:2, 4:3, 5:2, 5:3, 6:5, etc.), the patterns will consist of some loops (i.e., there will be intersecting lines), but there will also be uncompleted loops. This latter type of double-image pattern is represented by the 3:2 Lissajous figure of Fig. 12-54B. Note that this latter image has one completed loop and two "free" or dangling ends. Actually these ends are not really "free," since in tracing the pattern, the spot arriving at one of these end points retraces its path in the opposite direction, thus completing the loops; however, the end points *appear* to be unconnected in each fractional double image. *All double-image patterns are characterized by two free ends.* The exact position of these two dangling ends is dependent upon the frequency ratio and the deflection plates to which the signals are applied.

Lissajous patterns that represent input signals with initial phase angles of 0°, 90°, 180°, or 270° are best to use in determining the frequency ratio. If the resultant picture is a double-image pattern, the initial phase can only be *one* of the above four values. If the pattern consists entirely of completed loops and is symmetrical about both the X- and Y-axes, then the input signals causing this pattern will have an initial phase angle of either *one or two* of the above four values.

For the remainder of this chapter, we will have occasion to study Lissajous patterns for many other frequency ratios, both of the symmetrical completed loops types and the double-image patterns.

Frequency Calculations from the Patterns

Now that we have a fair idea of the general types of Lissajous patterns and under what circumstances they are formed, it becomes a very simple matter to determine the frequency relationship between the two input sine-wave signals from the patterns. By drawing (or visualizing) a tangent to one vertical side of the pattern and another tangent to one horizontal side of the pattern (it does not matter to which side the tangents are drawn), it is possible to determine the frequency ratio. The frequency ratio can be expressed in simple mathematical form as follows

$$\frac{f_v}{f_h} = \frac{t_h}{t_v} \qquad (12\text{-}4)$$

where f_h = frequency of horizontal-deflection voltage
f_v = frequency of vertical-deflection voltage ,

t_h = number of loops which touch the horizontal tangent line

t_v = number of loops which touch the vertical tangent line.

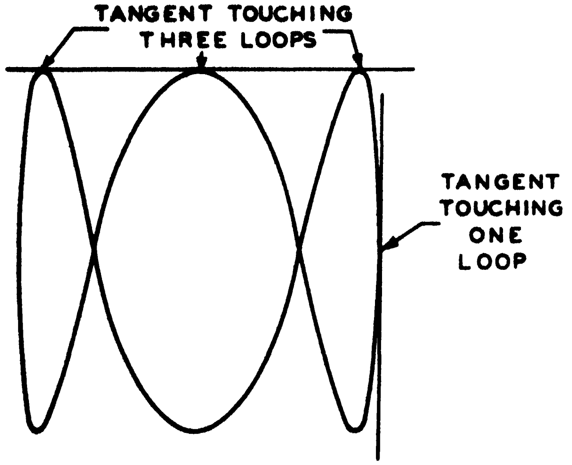

Fig. 12-55.—Method for calculating an unknown frequency by Lissajous figures.

For instance, let us examine the pattern of Fig. 12-55, which is the Lissajous figure of Fig. 12-48C redrawn with vertical and horizontal tangent lines. From this drawing we find that t_h is equal to 3 and t_v is equal to 1, thereby establishing the frequency ratio as 3:1. This frequency ratio is also evident from the completed loop patterns (B) and (D) of Fig. 12-51. You can easily visualize the pattern of Fig. 12-53C as having three horizontal tangent points to two vertical tangent points, which tells us that the frequency ratio of f_v/f_h is 3:2. The t_h/t_v ratio of Fig. 12-52B, C, or D is seen to be 1:3, rather than 3:1, which means that the frequency of the horizontal-deflection signal is three times as great as that of the vertical-deflection signal.

Thus far we have seen how to interpret the frequency ratio from completed loop patterns. When dealing with double-image patterns, the situation is somewhat different. We can still utilize the formula of equation (12-4), but where a tangent line touches an uncompleted loop (a free end) it should be counted as ½ a tangent point, and completed loops as a single point. For example, refer to Fig. 12-56, which indicates a number of double-image patterns with tangent lines drawn. In pattern (A), we find that t_h is equal to 1 and t_v equal to ½, which means that the t_h/t_v ratio is 1/½, or 2:1. This establishes the vertical-deflection voltage as having a frequency equal to twice that of the horizontal-deflection voltage. If a lower horizontal tangent line of Fig. 12-56A were used instead, it would touch two free ends, and so $t_h = \frac{1}{2} + \frac{1}{2}$ or 1, the same as before.

For pattern (B) of Fig. 12-56, t_h is equal to 1½ and t_v equal to $\frac{1}{2} + \frac{1}{2}$ or 1, so that t_h/t_v is 1½/1 or 3/2, which means that the frequency of the vertical-deflection voltage is one and one-half times as great as that of the horizontal-deflection voltage. In part (C) of Fig. 12-56, the t_h/t_v ratio is found to be 1½/½ which is equal to $\frac{3}{2}/\frac{1}{2}$, or a 3:1 ratio. Thus we find that f_v/f_h for this pattern is 3:1.

The determination of greater frequency ratios represented by other Lissajous patterns is carried out in the same manner just described. Although the patterns are more complex in appearance, the same method still applies In the previous examples, sine waves were used as the input signals because frequency standards invariably produce sine-wave signals and also because the resulting Lissajous patterns are easier to interpret. The Lissajous patterns discussed in this chapter will be based on sine-wave inputs.

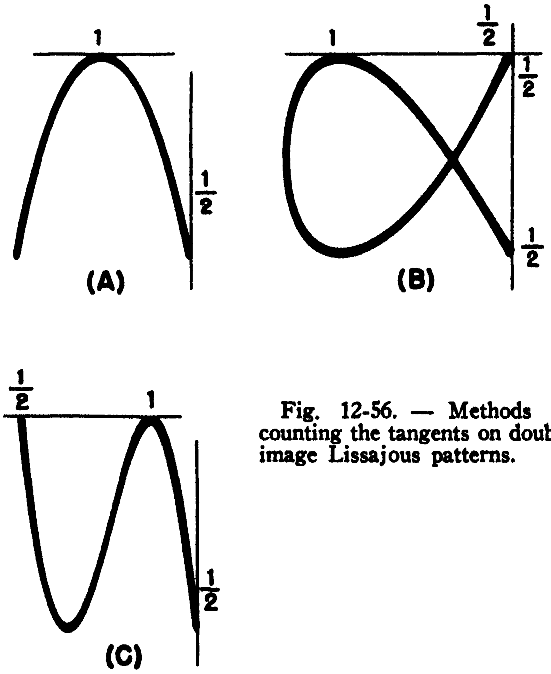

Fig. 12-56. — Methods for counting the tangents on double-image Lissajous patterns.

Frequency Calibration and Frequency Standards

By the use of Lissajous figures, we have a ready means of comparing a known frequency against an unknown, and hence can determine the frequency of the latter. This can be carried out to a high degree of accuracy. In the use of the oscilloscope for this purpose, it is important to know the frequency response of the amplifiers in the oscilloscope. In other words, if the amplifiers of the oscilloscope are to be used, the frequency of the input signals must fall within the frequency response of these amplifiers, so that both signals appear at their respective deflection plates. (The frequency responses of practically all the popular commercial cathode-ray oscilloscopes are given in Chapter 22.)

Not only are Lissajous patterns used for determining the frequency of an unknown signal, but more generally they may be used as a means of calibrating one instrument against a standard. In the comparison of the frequencies or the calibration of some equipment, one frequency must be known and the other is unknown. The frequency standard should be an accurate oscillator of any type. The range of the oscillator used as a standard is not so important, provided that its highest known frequency is at least *one-tenth* of the highest frequency to be calibrated upon the unknown. Thus, with a *single* 1,000-cps standard, it is possible to calibrate a range of from 100 to 10,000 cps. If the standard is continuously variable and has a range fully as great as the expected range of the unknown or uncalibrated oscillator, calibration is, of course, greatly simplified.

Also, it must known whether the standard frequency voltage has been applied to the vertical plates or to the horizontal plates. If the standard frequency is higher than the unknown and is applied to the vertical plates, the loops on the *top* (or bottom) of the pattern will be greater than the number of loops on either *side* of the pattern. Of course, when a 1:1 ratio is reached, the image will be a circle, ellipse, or straight line. If the standard frequency is the lower frequency and if it is held constant at its original value and the unknown is advanced, the greater number of loops will appear on the side of the pattern and the smaller number of loops will appear on the top of the pattern. Because of this change in the number of loops on the side or top of the pattern, it is necessary to know which plates are being supplied with the standard voltage.

Another important consideration, when working with *completed-loop patterns,* is to make certain that all of the loops in the image are shown and that none are covered by others. If the pattern cannot be stopped on the screen or if the pattern appears very complex, it is best to work with a pattern which is moving very slowly, so that all loops may be noticeable. Under certain conditions, the linear sweep circuit in the cathode-ray oscilloscope may be used as a standard, but only after certain frequency settings of this system have been definitely established and are accurately known.

Sine waves, nonsinusoidal waves, and modulated waves may be used during such frequency comparison, but not with equal facility. The greatest ease of interpretation, which means the most rapid and accurate determination, is obtained when two sine-wave voltages are compared.

Relative Amplitude of the Two Input Signals

Much can be said about the relative amplitudes of the two voltages applied to the vertical- and horizontal-deflection plates. However, efficient operation is best accomplished by having the image on the screen as large as possible, yet not too large (that is, not so large that the boundaries of the pattern extend to the limits of the screen). If the image is too large, the edges of the complex pattern will be hazy and interpretation will be difficult.

There is very little choice between a square-shaped pattern on the screen and a rectangular pattern; whichever suits the requirements best for interpretation should be used. The simpler the frequency ratio between the two voltages, the smaller may be the pattern. The more complex the frequency ratio between the two voltages, the larger must be the image, so as to enable visual inspection and proper interpretation. At the same time, it should also be realized that the larger the image, the less its intensity and the less distinct will be the lines, loops, and intersections, all of which may be involved in making the correct interpretation. Any effort to make the image brighter may result in impairment of the focus, thus making the boundaries of the image even more indistinct and difficult to interpret.

Other Lissajous Patterns

Ratios as high as 10:1 and as low as 10:9 can be observed with patterns similar to those shown in Figs. 12-51 and 12-52. In each case, tangents drawn to two sides, and counting the number of points (as previously explained) which touch these tangents, will identify the frequency ratio. Space does not permit a complete explanation with full data concerning the graphical development of more complex types of Lissajous figures. Development of the most intricate pattern progresses exactly as outlined for the simple patterns. However, the actual construction is naturally more detailed and more intricate, because the pattern becomes more complex at higher ratios, and a greater number of spot positions must be established in order to construct the pattern. While it is true that the pattern resulting from, for example, a 3:2 ratio, obtained by applying two sine voltages, will differ from that developed when one of these voltages is a sawtooth wave, the actual method of development remains the same, as was indicated in Chapter 9 on Linear Time Bases.

If you are interested in graphically solving other Lissajous patterns of the general ratio *a/b,* draw *a* cycles for the vertical-deflection voltage and *b* cycles for the horizontal-deflection voltage, and, employing the same time interval for both, mark off the respective amplitudes and project the resultant spot displacement, as has been done for the previous graphical illustrations in this chapter. Remember that the lower the ratio (i.e., the closer the fractional ratio approaches unity, such as 6:5, 8:7, 10:9, etc.), the more complex the pattern becomes and hence the more points will be needed to construct the pattern.

Thus far we have studied various 1:1, 2:1, 3:1, and 3:2 ratio patterns by means of Lissajous figures employing sine-wave input signals. In order to indicate how the Lissajous patterns would appear when one of the input signals is a modulated sine wave, refer to Fig. 12-57. Photos (A) and (B) both indicate 2:1 frequency ratios and (C) a 3:1 ratio. For pattern (A), a 100-kc sine wave, amplitude-modulated by a 400-cps sine wave, is the input signal to the vertical-deflection plates, and a 50-kc sine wave is the input to the horizontal-deflection plate. For pattern (B), a 100-kc sine wave amplitude-modulated at 400 cps is the input to the horizontal-deflection plates. The degree of the modulation for the input signal of pattern (A) is greater than that for pattern (B). From both patterns, the 2:1 frequency ratio is easily seen. For the pattern at (C), the vertical-deflection voltage is a 150-kc sine wave, amplitude-modulated by a 400-cps sine wave, and the horizontal-deflection voltage is a 50-kc sine wave, thus producing a 3:1 ratio pattern.

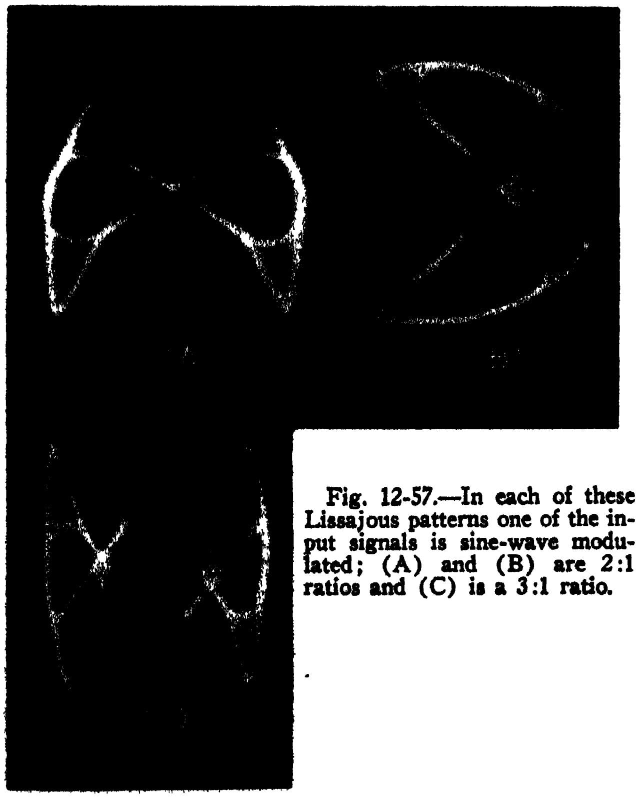

Fig. 12-57.—In each of these Lissajous patterns one of the input signals is sine-wave modulated; (A) and (B) are 2:1 ratios and (C) is a 3:1 ratio.

The following Lissajous patterns (Figs. 12-58 through 12-85) will indicate frequency ratios possible, up to 10:1, which have not yet been covered. No modulated signals will be employed. For each figure, part (A) represents the symmetrical completed-loop pattern and part (B) an uncompleted or double-image pattern with free ends. Both parts of each figure represent the same frequency ratio, the only difference between the two input signals residing in their initial phase condition. For all cases shown, the higher-frequency signal is that applied to the vertical-deflection plate. This means that in all patterns the greater number of loops or points will lie along the horizontal tangent line. To visualize how the pattern will appear if the input signals to the deflection plates are reversed, simply rotate the page by 90° in a clockwise direction.

The patterns will be presented in groups in the order of increasing numerator of the ratio. Patterns representing 4:1 and 4:3 ratios will fall in *group 4,* patterns indicating 5:1, 5:2, 5:3, and 5:4 will fall in *group 5,* etc. The ratio indicated by any of the patterns is a function of the relative frequencies of the two voltages. Regardless of the exact numerical value of these frequencies, the pattern will be the same for any one particular ratio. In other words, a 6,000-cps and a 1,000-cps voltage will create a 6:1 ratio pattern and this pattern will remain the same if the two frequencies are doubled to 12,000 and 2,000 cps, or changed to 12,600 cps and 2,100 cps, respectively, 120,000 and 20,000 cps, etc. In other words, the ratios will be reduced to their lowest common denominators.

Group 4 Ratios

Since we have covered, in detail, Lissajous patterns of 1:1, 2:1, 3:1, and 3:2 ratios, the next ratio will have 4 as the numerator. In group 4, only two possible ratios exist which have not been discussed. They are 4:1 and 4:3 ratios. (A 4:2 ratio is the same as a 2:1 ratio and a 4:4 ratio is the same as a 1:1 ratio, and both types have already been studied.)

Fig. 12-58A and B illustrate two patterns for a 4:1 ratio. The frequency relationship for either pattern can be calculated from equation (12-4). The 4:1 ratio is readily seen from the completed-loop pattern of part (A), but it takes a few seconds longer to visualize this ratio from pattern (B) because of the necessity of

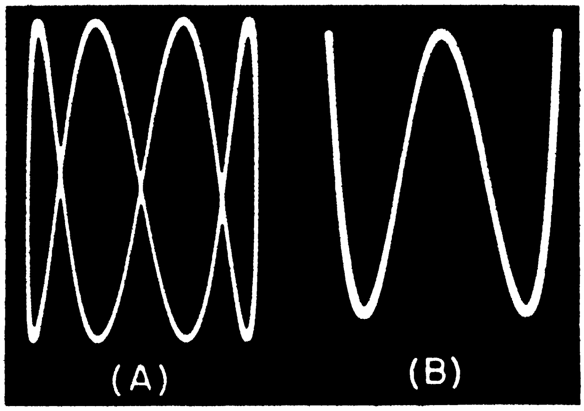

Fig. 12-58.—Oscillograms of 4:1 frequency ratios.

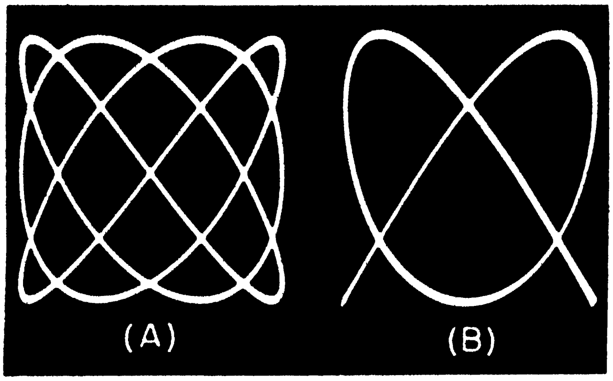

Fig. 12-59.—Oscillograms of 4:3 frequency ratios.

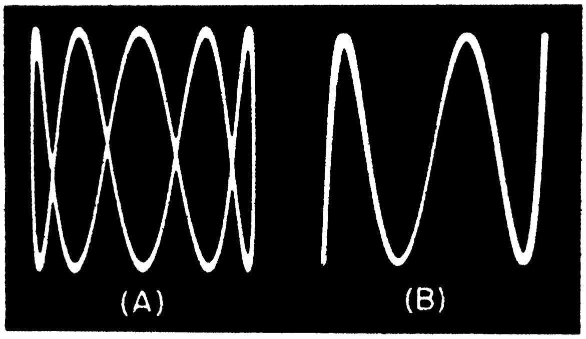

Fig. 12-60.—Oscillograms of 5:1 frequency ratios.

counting the free ends as only ½. At the lower ratios, the completed-loop type pattern is simpler to use, but for higher ratios, as will soon be evident, it is best to use the double-image type pattern. In Fig. 12-59A and B appear two Lissajous patterns, each depicting ratios of 4:3.

Group 5 Ratios

In this group, where the numerator is 5, four ratios are possible; they are 5:1, 5:2, 5:3, and 5:4. The two patterns of Fig. 12-60 indicate a 5:1 ratio. Note that the double-image pattern of part (B) does not have any intersecting lines. In this latter photo, t_h equals 2½ and t_v equals ½, making the f_v/f_h ratio 2½/½ or 5:1.

In Figs. 12-61, 12-62, and 12-63 appear the Lissajous patterns for ratios of 5:2, 5:3, and 5:4, respectively. Counting the loops or peaks along the horizontal of part (A) of each of these photos readily indicates the number to be five. Counting the completed loops and free ends of part (B) of these photos indicates that there are 2½ horizontal tangent points for each. For parts (A) of Figs. 12-61, 12-62, and 12-63, the loops or peaks along a vertical side increase from 2 to 3 to 4, respectively, while for parts (B), we count the vertical points as 1, 1½, and 2, respectively.

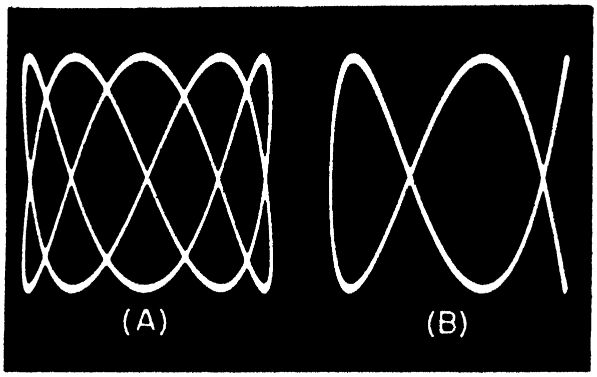

Fig. 12-61.—Oscillograms of 5:2 frequency ratios.

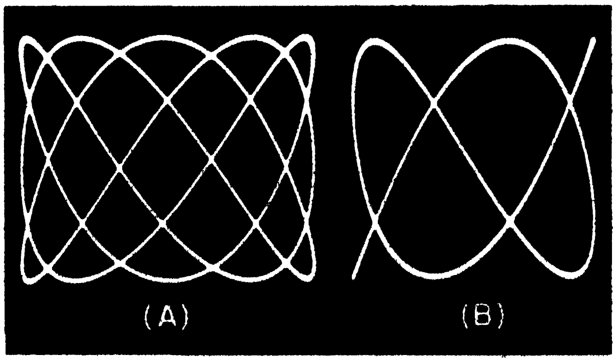

Fig. 12-62.—Oscillograms of 5:3 frequency ratios.

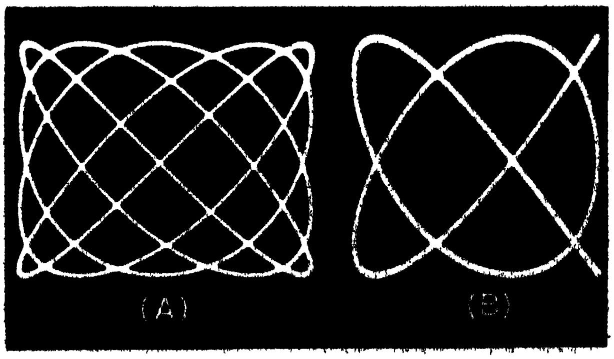

Fig. 12-63.—Oscillograms of 5:4 frequency ratios.

Group 6 Ratios

In group 6, only two new ratios are possible; they are 6:1 and 6:5, and they appear in Figs. 12-64 and 12-65, respectively. Note the six peaks on top and one on the sides of Fig. 12-64A, while for part (A) of Fig. 12-65, we count six peaks along the horizontal and five peaks along the vertical.

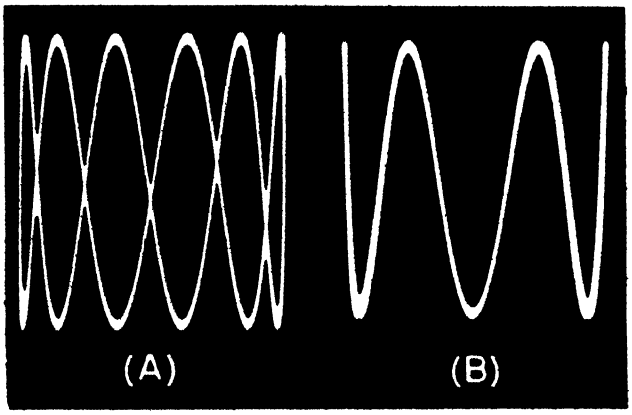

Fig. 12-64.—Oscillograms of 6:1 frequency ratios.

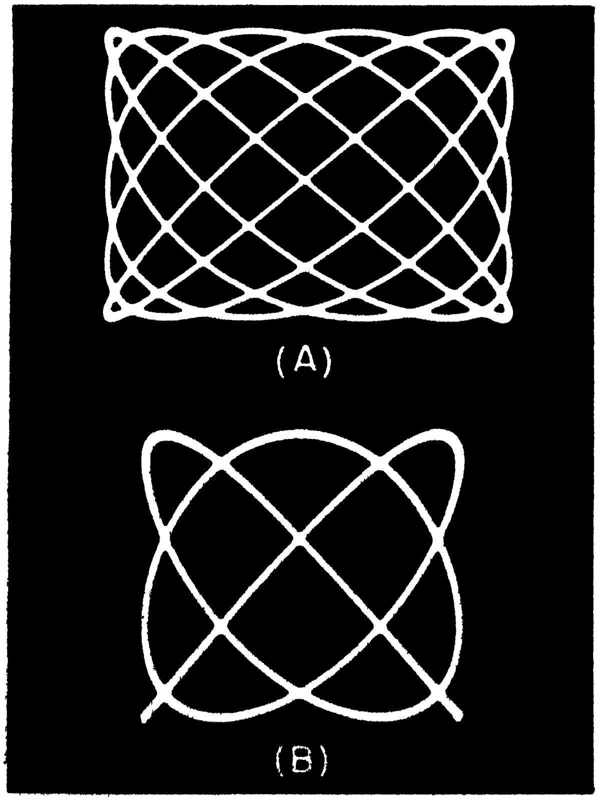

Fig. 12-65.—Oscillograms of 6:5 frequency ratios.

Group 7 Ratios

There are six new ratios possible in group 7, which are 7:1, 7:2, 7:3, 7:4, 7:5, and 7:6. The Lissajous patterns for these ratios appear in Figs. 12-66 through 12-71, respectively. In each of the completed-loop patterns of part (A) of these drawings, we find seven peaks lying along the horizontal. Therefore, for each pattern, we consider the numerator of the ratio as seven. The vertical sides of patterns (A) indicate that each of their peaks successively increases by one, so that there are from one to six as we progress from Fig. 12-66 through Fig. 12-71. The horizontal of parts (B) indicates 3½ tangent points, while along the vertical, the tangent points increase from ½ to 3, in steps of ½, for Figs. 12-66 through 12-71.

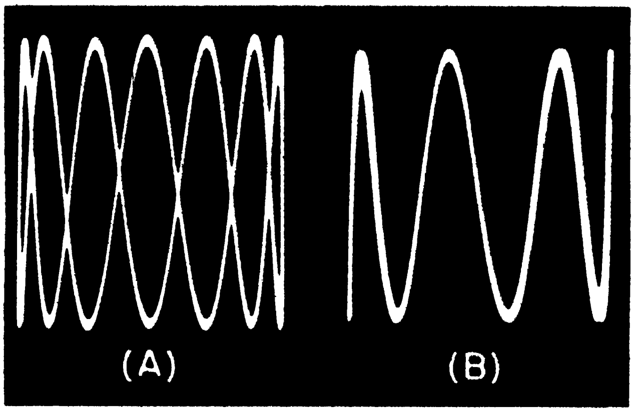

Fig. 12-66.—Oscillograms of 7:1 frequency ratios.

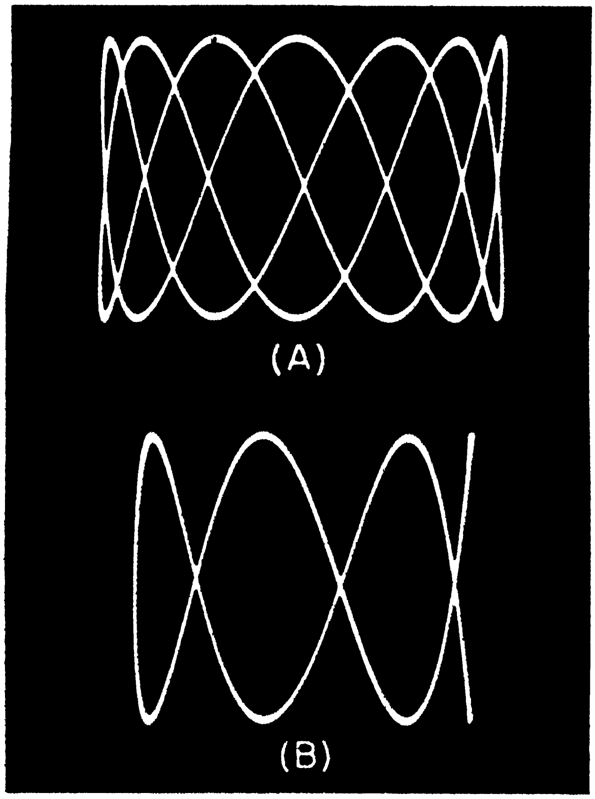

Fig. 12-67.—Oscillograms of 7:2 frequency ratios.

An interesting thing about the double-image patterns, parts (B), is the position assumed by the free

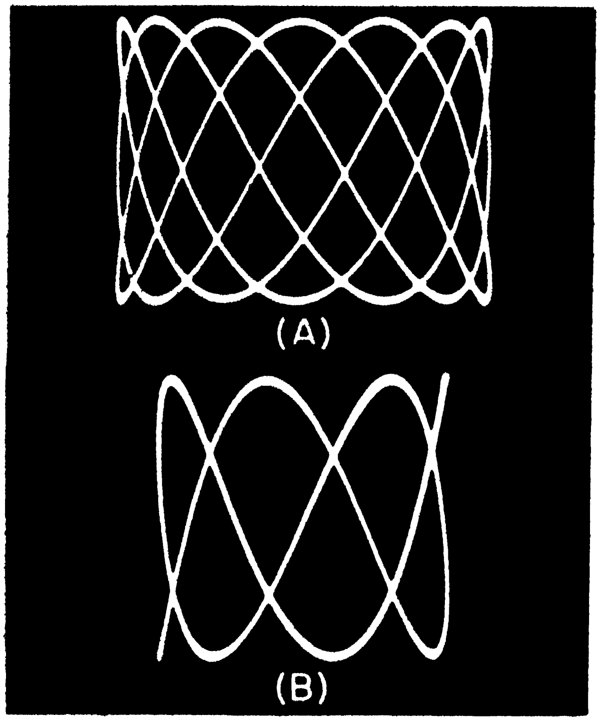

Fig. 12-68.—Oscillograms of 7 :3 frequency ratios.

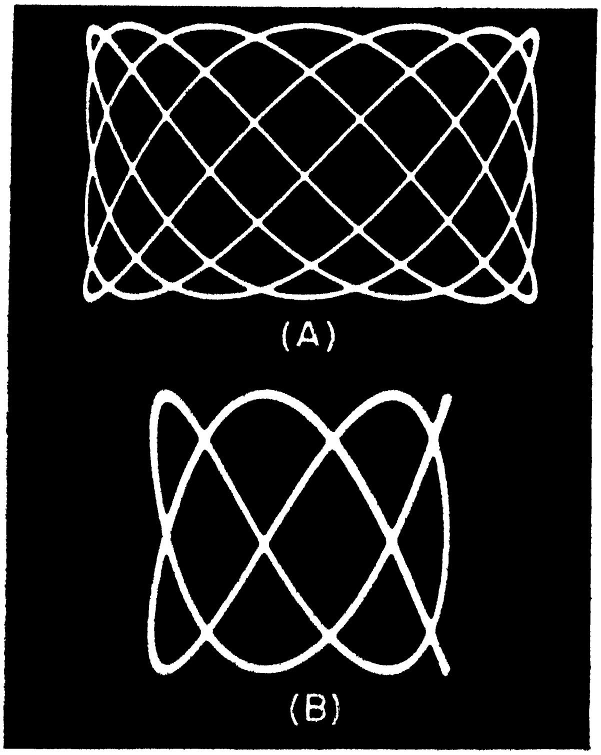

Fig. 12-69.—Oscillograms of 7 :4 frequency ratios.

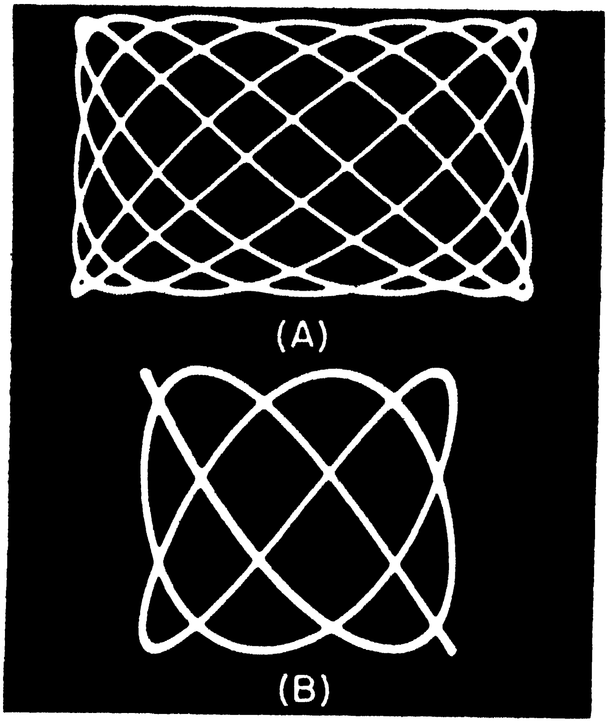

Fig. 12-70.—Oscillograms of 7 :5 frequency ratios.

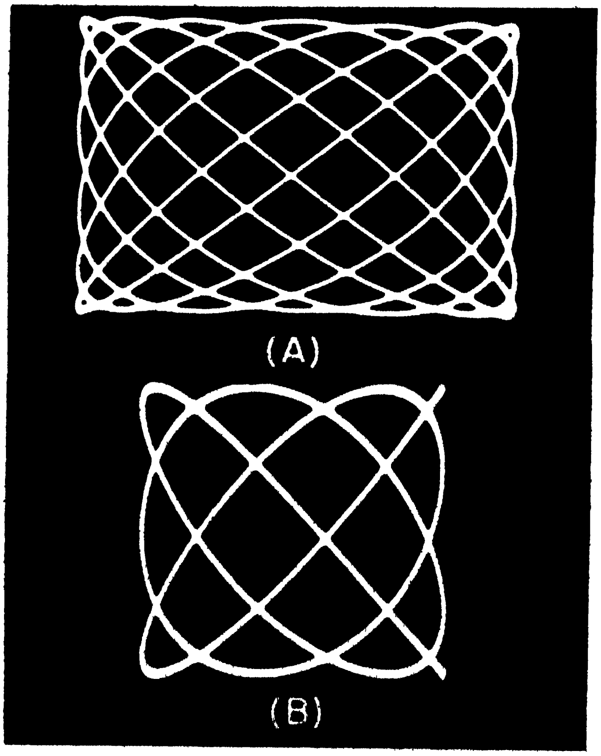

Fig. 12-71.—Oscillograms of 7 :6 frequency ratios.

ends, that is, those parts which we count as ½ a tangent point. In those double-image patterns where the denominator of the ratio is *even*, that is, for ratios 7 :2, 7 :4, and 7 :6, both free ends lie along the same side. In Figs. 12-67, 12-69, and 12-71, this is seen to be along the right-hand vertical side. Where the demoninator of the ratio is odd, as indicated by ratios 7 :1, 7 :3, and 7 :5, one free end will be *diagonally* opposite the other, as seen in Figs. 12-66, 12-68, and 12-70. This same effect exists also in all other cases where the numerator of the ratio is itself an *odd* number. If the numerator is an *even* number, then the situation concerning even- and odd-numbered denominators would be reversed.

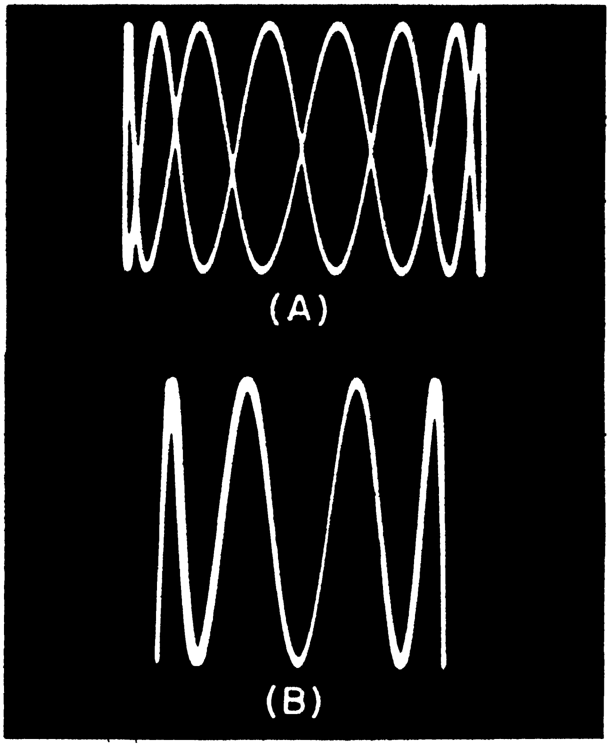

Fig. 12-72.—Oscillograms of 8:1 frequency ratios.

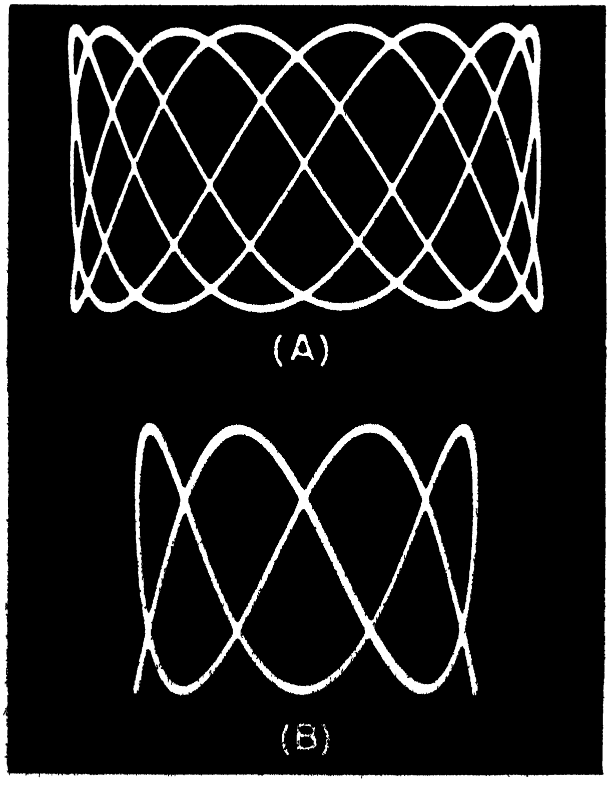

Fig. 12-73.—Oscillograms of 8:3 frequency ratios.

Group 8 Ratios

Only ratios 8:1, 8:3, 8:5, and 8:7 are possible in this group. In Figs. 12-72, 12-73, 12-74, and 12-75

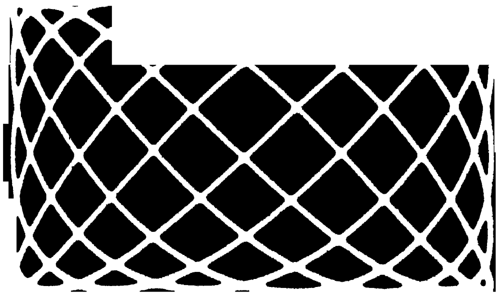

Fig. 12-74.—Oscillograms of 8:5 frequency ratios.

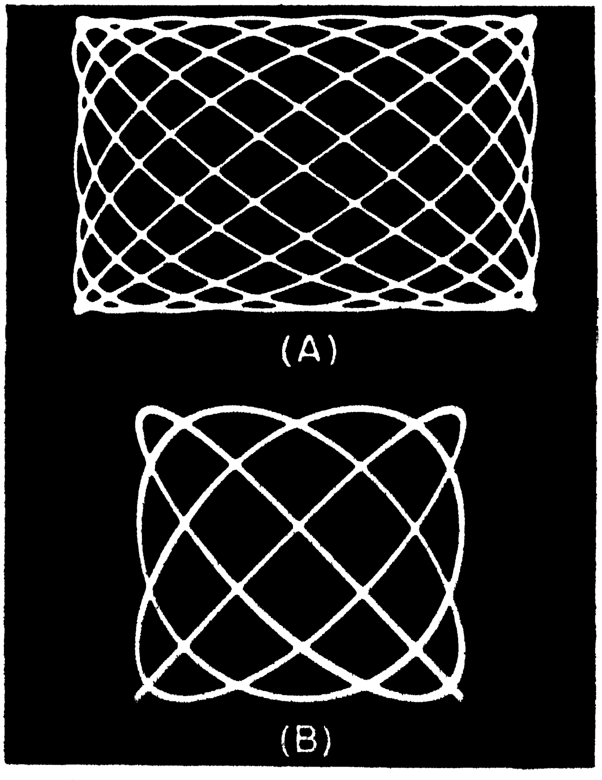

Fig. 12-75.—Oscillograms of 8:7 frequency ratios.

appear the Lissajous patterns for frequency ratios 8:1, 8:3, 8:5, and 8:7, respectively.

You will note from Fig. 12-75, which indicates a ratio, 8:7, approaching unity, and where the numerator and denominator are both high, that it is much easier in these cases to determine the frequency ratio from the double-image pattern of part (B) than from the completed-loop pattern (A).

Group 9 Ratios

In this group, six different ratios are possible. These ratios are 9:1, 9:2, 9:4, 9:5, 9:7, and 9:8, and are shown in Figs. 12-76 through 12-81, respectively. Nine peaks appear along the horizontal for each part (A) of the photos. As the ratios decrease, that is, as we approach 9:7 and 9:8 ratios, it becomes difficult to count the exact number of peaks along the horizontal and vertical of the completed-loop patterns. At these ratios, it is better to use double-image patterns for the determination of the ratio because of the clarity of the image. You can readily see how much easier it is to interpret the frequency ratio of 9:7 and 9:8 from photos (B) of Figs. 12-80 and 12-81, respectively, than from the photos of part (A) of the same figures.

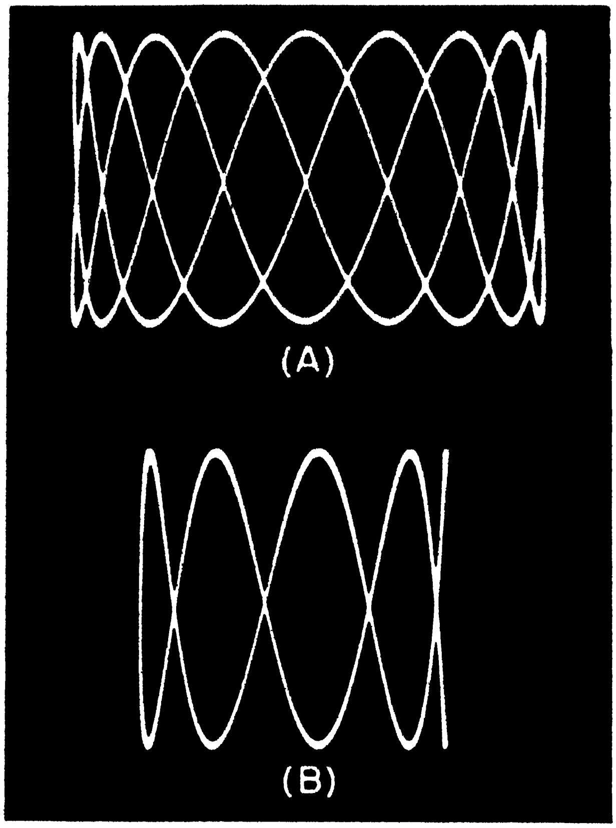

Fig. 12-77.—Oscillograms of 9:2 frequency ratios.

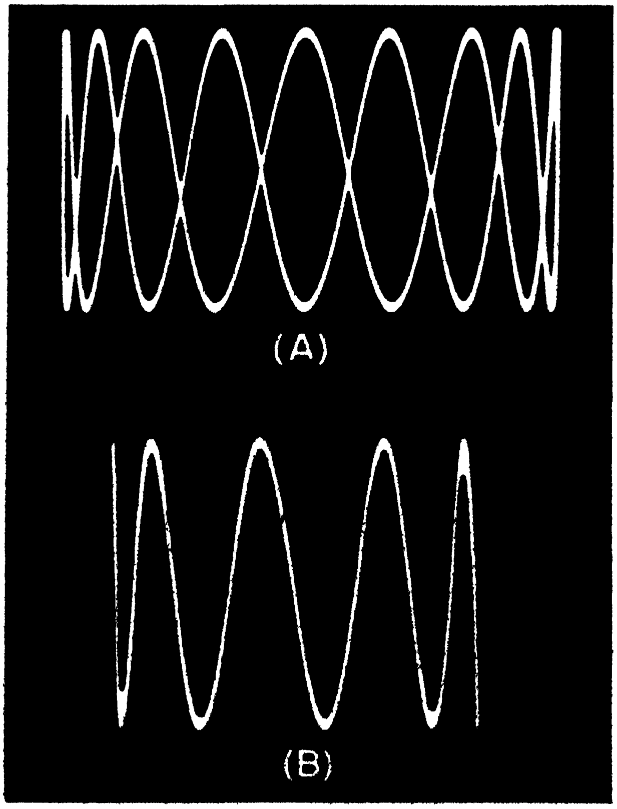

Fig. 12-76.—Oscillograms of 9:1 frequency ratios.

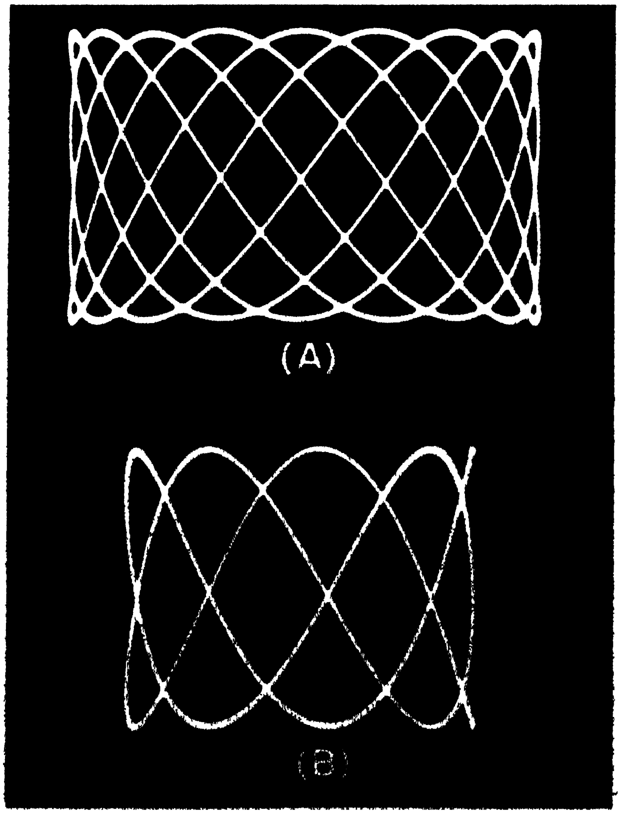

Fig. 12-78.—Oscillograms of 9:4 frequency ratios.

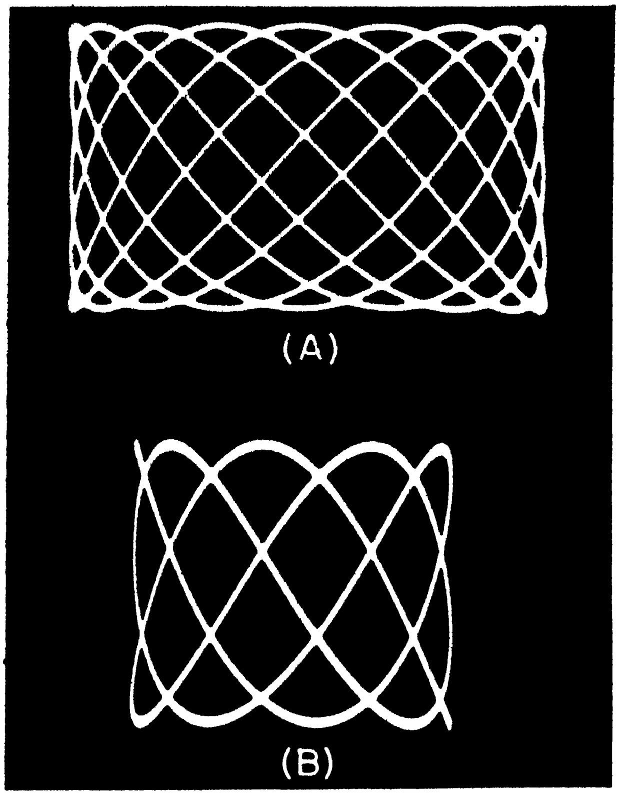

Fig. 12-79.—Oscillograms of 9 :5 frequency ratios.

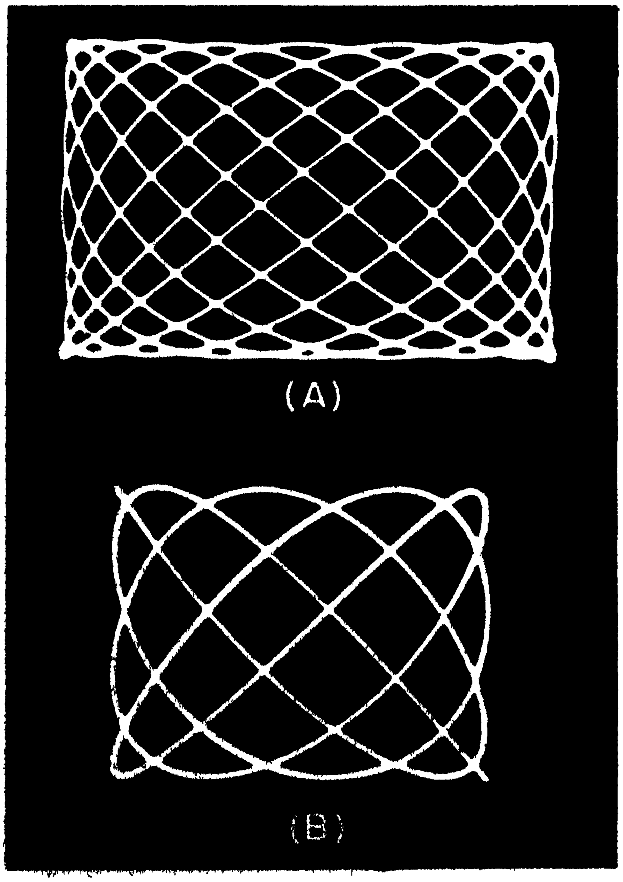

Fig. 12-80.—Oscillograms of 9 :7 frequency ratios.

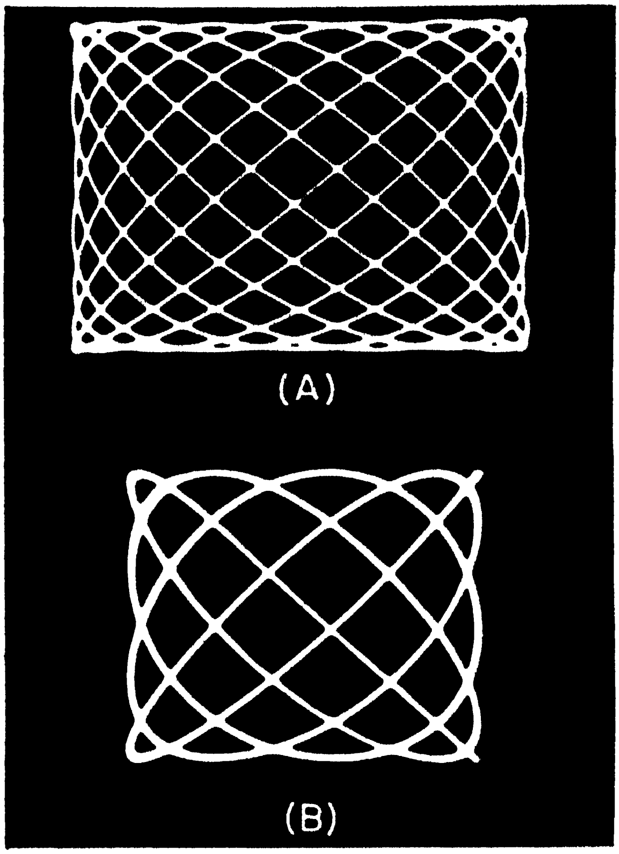

Fig. 12-81.—Oscillograms of 9 :8 frequency ratios.

Group 10 Ratios

In this last group we are to study, only four ratios are possible. These ratios are 10 :1, 10 :3, 10 :7, and 10 :9. The completed-loop and double-image patterns representative of these ratios appear in Figs. 12-82, 12-83, 12-84, and 12-85, respectively. From examination of the completed-loop patterns (A) of Figs. 12-84 and 12-85, we find that it is extremely difficult to determine the frequency ratios from these Lissajous figures. On the other hand, we experience no difficulty in calculating these same frequency ratios from the double-image patterns (B) of Figs. 12-84 and 12-85.

In the Lissajous patterns of Figs. 12-58 through 12-85, the lowest input frequency used was 500 cps and the highest frequency was 5,000 cps.

From the analysis of Lissajous figures, we see that there is a limitation as to how far we can use this method of frequency comparison, especially as the ratios approach unity. This is so because the patterns become difficult to interpret. There are other methods of frequency comparison, although not as popular as the Lissajous pattern method, which can be used to interpret higher ratios and those closer to unity. Of course, these other methods can also be used for the simpler frequency ratios. In the following sections, we will discuss a number of these methods, show some typical oscillogram patterns, and explain how they are formed.

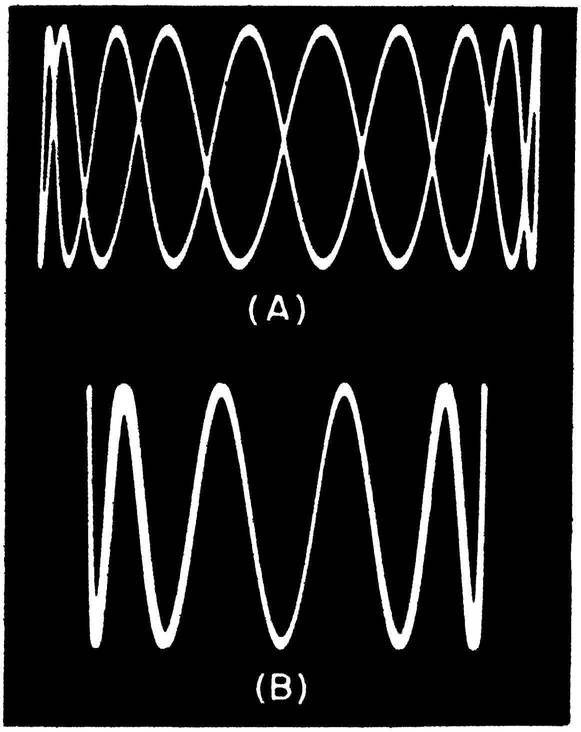

Fig. 12-82.—Oscillograms of 10 :1 frequency ratios.

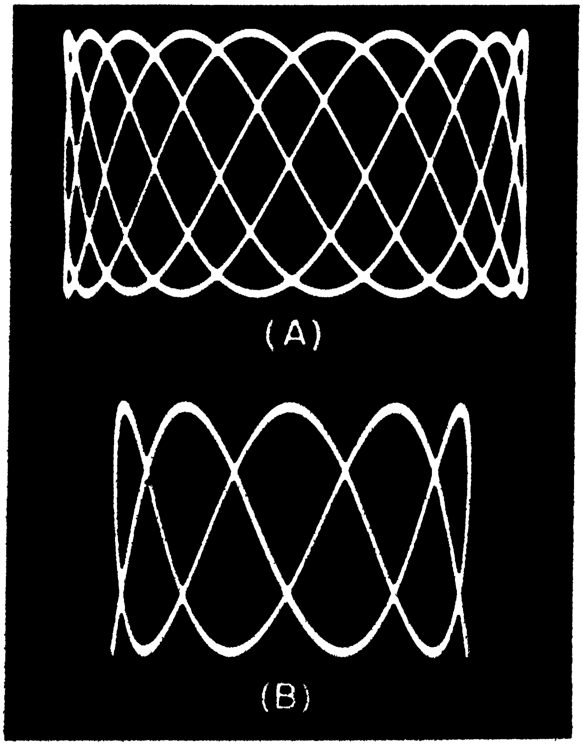

Fig. 12-83.—Oscillograms of 10 :3 frequency ratios.

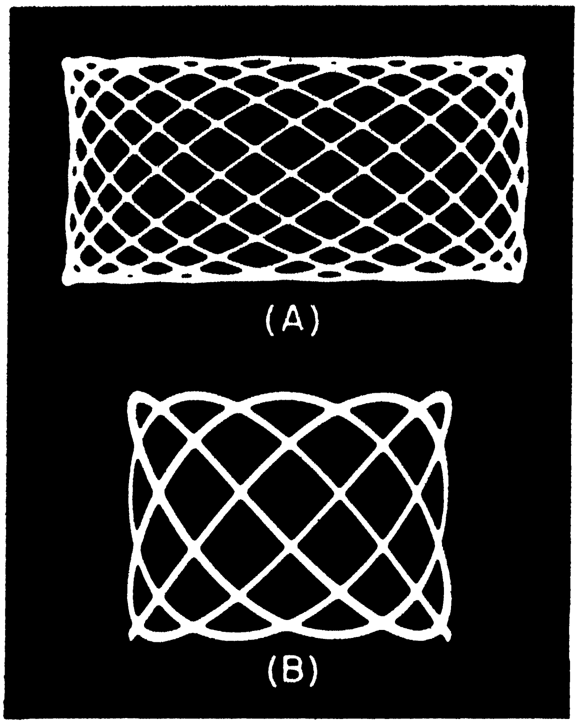

Fig. 12-84.—Oscillograms of 10 :7 frequency ratios.

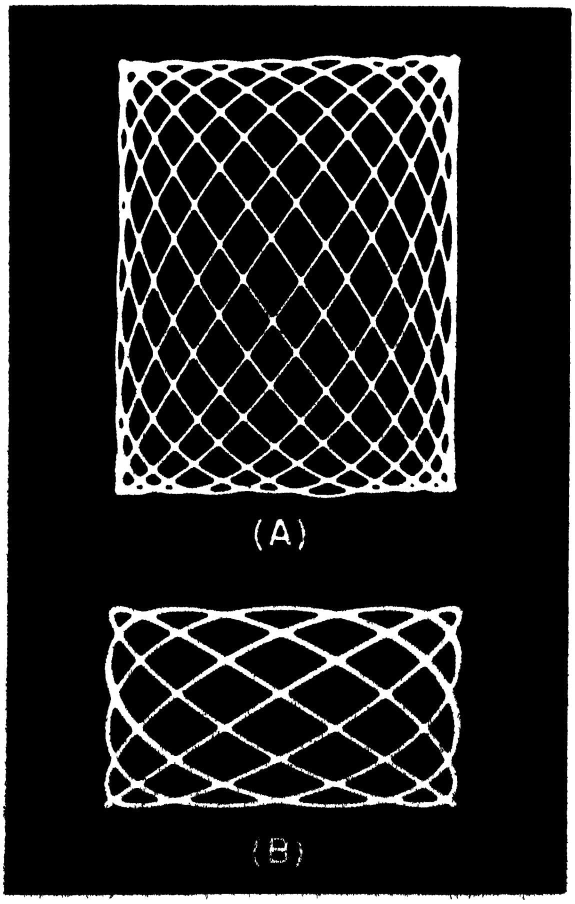

Fig. 12-85.—Oscillograms of 10 :9 frequency ratios.

Intensity Modulation

In the section of this chapter where we discussed a method of determining the direction of spot travel, we showed how the pattern appears when intensity modulation is applied. In this section, we will show how, by the use of circular patterns, we can determine the frequency of an unknown signal for low and high ratios. These ratios are, however, limited to the simple cases where the frequencies of the two signals are integral multiples of each other. The chief advantage of this method, however, is that the unknown signal can be almost any periodic waveshape without making the resultant pattern difficult to interpret.

We already know that for a 1:1 Lissajous figure where the phase difference is 90° or 270°, a circular pattern will result if the amplitudes of the two input signals are equal, while a symmetrical elliptical pattern will be produced when the amplitudes are different. If we take a sine-wave signal and pass it through a phase-splitting network, which is properly connected to the oscilloscope, we can obtain this circular or elliptical pattern by this single signal alone. If we apply the other signal, which is to be compared to the first one, to the grid circuit of the cathode-ray tube we will intensity-modulate the electron beam. If this latter signal is strong enough, it will cause parts of the trace to be blanked out during its negative half-cycles. The number of blanking intervals or trace periods as seen on the screen determines the ratio of the two signals. We indicated this in Fig. 12-36 where the ratio was 10:1. Before we illustrate some typical patterns, let us briefly describe how a simple phase-splitting network produces the circular or elliptical pattern. A simple and very common type of phase-splitting network consists of a capacitor and resistor in series as indicated in Fig. 12-86A.[6] The voltage drop E_R across the resistor is equal to the series current I flowing in the circuit multiplied by the value of resistance (by Ohm's law) and the voltage drop across the capacitor is equal to the current I multiplied by the capacitive reactance of C.

It is a well-known fact that the current flowing through a (pure) resistor is in phase with the voltage drop across it. This in-phase condition is illustrated in the vector diagram in part (B) of the figure, where the current vector I is seen to be in phase with the resistor voltage vector E_R. It is also known that the current flowing through a (pure) capacitance *leads* the voltage across it by 90°. This phase relation is also shown in the vector diagram where the capacitor voltage vector E_C is seen to be 90° behind or lagging the current vector I. From this simple analysis, we find, from the vector diagram, that the resistive voltage drop E_R and the capacitive voltage drop E_C are 90° out of phase with E_R leading E_C.

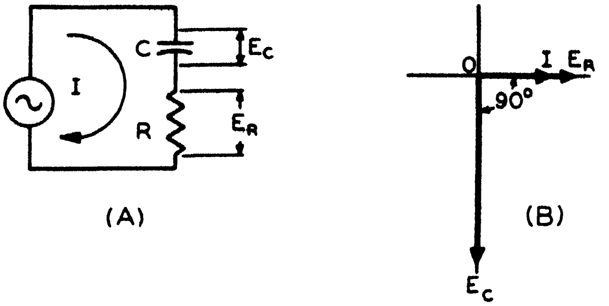

Fig. 12-86.—A single phase-splitting network is shown at (A), the vector voltage and current relations are indicated in (B).

For the sake of illustration, the length of vector E_C was made twice the length of vector E_R, which means that the capacitive reactance is twice as high as the resistance for this case.

In Fig. 12-87 are illustrated three schematic representations of connecting such an R-C phase-splitting network to the oscilloscope. It does not matter whether the signal is fed to the amplifiers of the oscilloscope (as shown) or directly to the deflection plates of the cathode-ray tube. If it is fed directly to the plates, however, the signal must be strong enough to cause noticeable deflection. The input terminals in both drawings are not designated as being vertical or horizontal because the voltages can be applied across either set of plates. The resistors are made variable so that the shape of the image can be adjusted to either an elliptical or circular pattern. The latter type of pattern is usually preferable and is the one used in this section.

The circuit of part (A) of Fig. 12-87 is employed when the input signal is obtained from a balanced system. From this schematic, it is easily seen that the voltage drops across the R and C elements are each effectively applied to one set of deflection plates. When working with a balanced output generator, care must be taken that the center-tap lead of the output circuit does not pick up any hum. This center-tap lead must *not* be grounded to the grounded terminals of the oscilloscope, as this will prevent the circuit from operating properly. If hum pickup does cause trouble, then the auxiliary circuit of Fig. 12-87B can be used. The coil represents the balanced output circuit from the gener-

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[6] Fig. 12-86 was obtained from the text "Understanding Vectors and Phase," by Rider and Uslan, published by John F. Rider Publisher, Inc., N. Y.

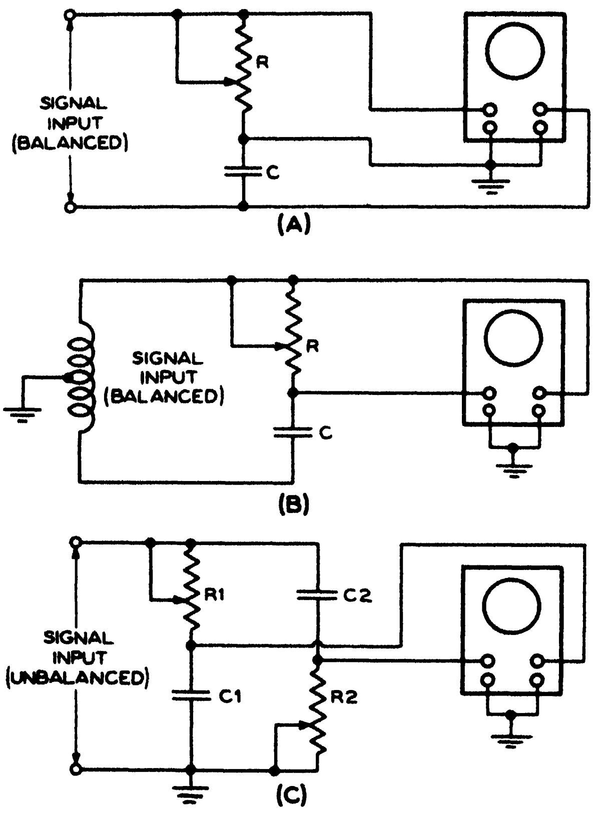

Fig. 12-87.—Three different schematic representation of how a phase-splitting network may be connected to an oscilloscope to produce a circular trace from a single input signal.

ator with the center tap grounded. The common connection between the resistor and capacitor can be connected to either ungrounded input terminal, and the other end of the resistor (as shown in the drawing) *or* capacitor is connected to the other input terminal.

When the signal source is obtained from a system that has an unbalanced output (one output terminal grounded) the circuit of part (C) of Fig. 12-87 should be used. From this drawing we see that two sets of phase splitters are employed. The resistances and reactances of the circuit components are all the same so that the voltage drops across any combination of resistor and capacitor are 90° out of phase with each other. (If *R1* is interchanged with *C1* and *R2* interchanged with *C2*, the circuit will still operate properly.) Therefore, by joining the common junction of each phase splitter to any one input terminal, a circular trace will result.

If both the unknown and known signal are sinusoidal, it does not matter which signal is used as the input to the phase-splitting network. If the unknown signal has a different waveshape, then it is best to use a known sinusoidal signal as the input to the phase splitter and the unknown as the input to the grid of the cathode-ray tube. Whatever the case may be, the known frequency should be so adjusted that the input signal to the phase-splitting network is the one of lower frequency. Let us now examine a few such intensity-modulated patterns.

In Fig. 12-88 appear two circular patterns which have been intensity-modulated. The frequency ratio is easily seen to be 4:1 by counting four trace or four blanking periods. These images, sometimes called *spot-wheel* patterns, were obtained using a 1,000-cps sine wave as the input signal to the phase-splitting network. Consequently, the frequency of the signal applied to the grid is 4 x 1,000 or 4,000 cps. Let us analyze why the ratio is calculated in this manner. The spot travels in the circular motion at a specific rate determined by the frequency of the signal applied to the deflection plates. A 1,000-cps signal causes the spot to travel or trace out the circle 1,000 times a second. This means that the spot will trace out one revolution or one circle in 0.001 second. If the signal applied to the grid has a frequency of 4,000 cps, it means that the beam will be blanked out once during one portion of each cycle. Since one cycle of a 4,000-cps signal occurs in 1/4,000 second or 0.00025 second, then in 0.001 second, which is the time for one complete revolution of the spot, *four* blanking periods will occur because $\frac{0.001}{0.00025} = 4$. Consequently, the frequency ratio is 4:1.

The signal applied to the grid of Fig. 12-88A was a 4,000-cps *sine* wave and that for Fig. 12-88B was a 4,000-cps *square* wave. Comparison of the trace sections of these two patterns can tell us which was modulated with the sine wave and which with the square wave. In the case of sine-wave modulation, the trace parts of the pattern vary in intensity, because the amplitude of the sine wave is not constant. This means that the individual traces will be more intense at their center because the sine wave will be a positive peak at

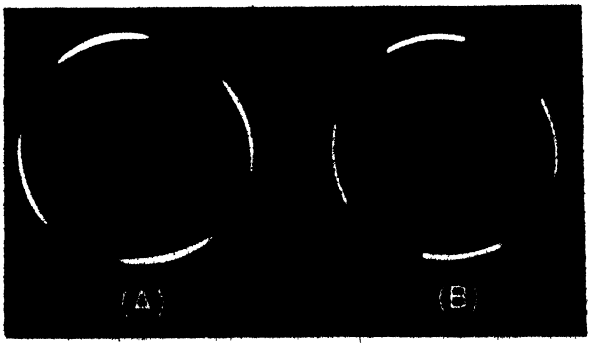

Fig. 12-88.—Frequency ratios of 4:1 are indicated by these intensity-modulated circular traces, (A) modulated by a sine-wave and (B) by a square wave.

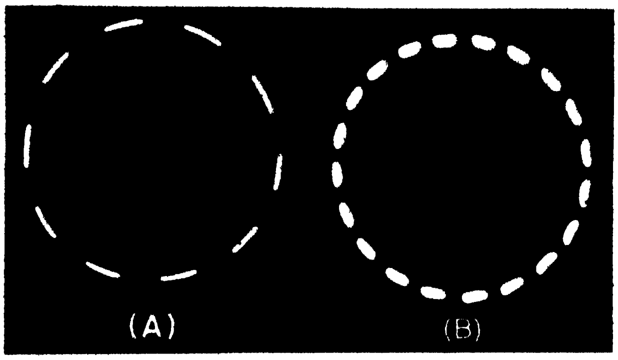

Fig. 12-89.—Intensity-modulated circular traces, (A) indicates a 10:1 frequency ratio and (B) a 20:1 ratio.

this time, and that the intensity of the traces will decrease evenly toward both ends. For the pattern which is modulated by the square wave, the intensity of the trace is constant because the amplitude of the square wave is constant during the trace periods.

In Fig. 12-89 appear two other spot-wheel patterns. Photo (A), which is similar to Fig. 12-36, is easily seen to indicate a 10:1 ratio. The frequency of the signal causing the circular pattern was a 500-cps sine wave and that modulating the grid was a 5,000-cps square wave. For photo (B), we count 20 traces indicating a ratio of 20:1. The signal causing the circular image in this latter case is a 500-cps sine wave, and that intensity-modulating the grid a 10,000-cps sine wave. From this latter photo, we see how much easier it is to interpret higher-frequency ratios by this method.

Other Frequency-Comparison Methods Using Phase-Splitting Circuits

In order to facilitate observations of higher ratios, it is possible to use the same phase-splitting network as in the previous example, but instead of modulating the grid, other electrodes are used. Fig. 12-90 illustrates one of these methods. The lower-frequency input is applied across the *R*-*C* phase-splitting network and the voltage drop across each element is applied to alternate sets of input terminals and hence to the deflection plates. The higher-frequency signal is applied to either input terminal through a resistor *R1* as seen in Fig. 12-90 and will modulate the circular or elliptical pattern.

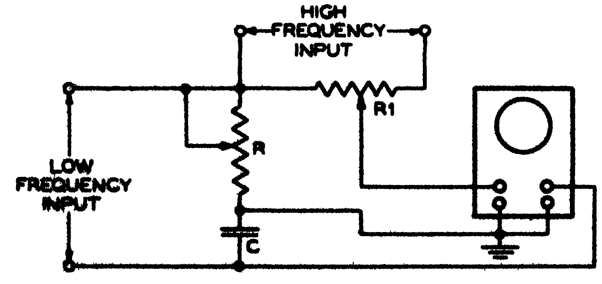

Fig. 12-90.—Another method of using a phase-splitting network for frequency comparisons.

Both signals should be of *sinusoidal* character. The higher-frequency signal will be displaced around the circle or ellipse formed by the lower-frequency signal. Resistor *R* is made variable so that either a circle or ellipse can be used. Resistor *R1* is made variable so as to control the amplitude of this modulating signal.

Previously we mentioned that Lissajous figures can be visualized as a wave traveling around a glass cylinder through which the front and rear patterns are visible. From the circuit of Fig. 12-90, the resulting patterns appear as though both these front and rear portions were separated and individually distinguishable. In other words, this method, irrespective of the initial phase difference between the two input signals, prevents any one portion of the pattern from being covered or coinciding with other parts. Let us now examine some patterns obtained by this method.

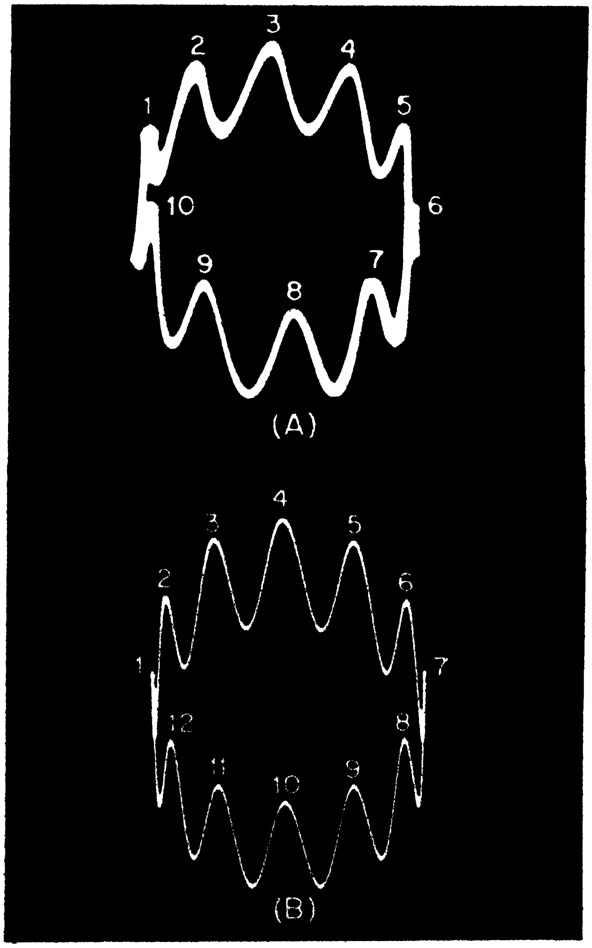

Fig. 12-91.—In these patterns part (A) indicates a 10:1 frequency ratio and (B) a 12:1 ratio.

In Fig. 12-91 appear two patterns where the higher-frequency signal is an integral multiple of the lower-frequency signal. Each pattern has its peaks numbered. Pattern (A) has 10 peaks and pattern (B) 12 peaks. This means that the numerators of the ratios are equal to 10 and 12, respectively. If the pattern appears as a single sine wave, that is, there are no intersecting lines (as in Fig. 12-91), then the denominator of the ratio is unity. Therefore, pattern (A) of Fig. 12-91 represents a 10:1 ratio and pattern (B) a 12:1 ratio.

In Fig. 12-92 appear two patterns where the denominator of the ratio is equal to two. By counting the number of peaks around the *top* of the pattern (visualizing the image as rotating around a cylinder) we immediately find out the numerator of the ratio. These peaks are numbered in the photo; for Fig. 12-92A, 13 peaks are counted, and 19 peaks appear for Fig. 12-92B. For the value of the denominator of the ratio, represented by the lower-frequency signal, we have to count the number of times that any part of the sine wave between a positive and negative peak is *intersected* and then add one to the result found. Thus, in Fig. 12-92, only one intersecting line is found for both patterns; hence, the denominator of the ratio for each pattern is 1 + 1 = 2. Therefore, we find that the ratio for pattern (A) is 13:2, and for pattern (B) it is 19:2. In Fig. 12-91, there were no intersecting lines, so the denominator is unity.

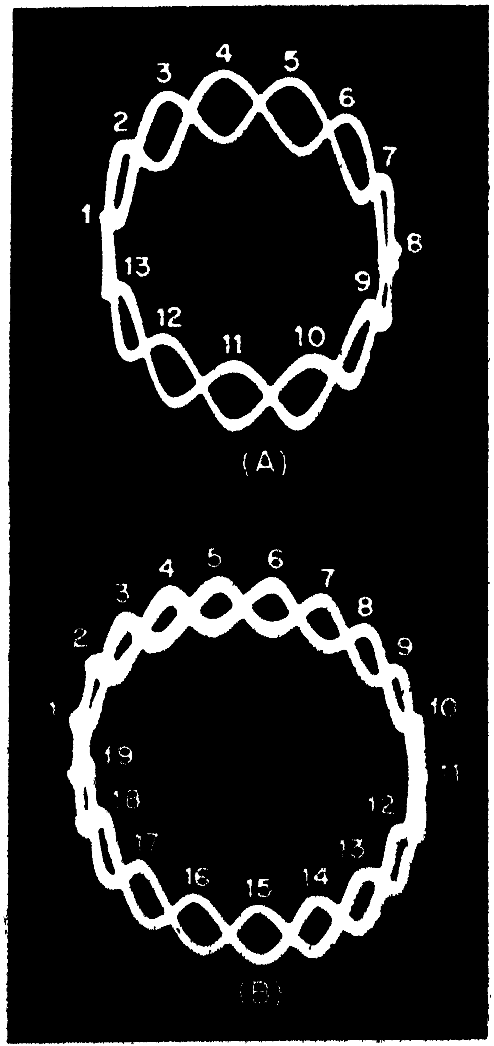

Fig. 12-92.—The oscillogram at (A) is for 13:2 frequency ratio and that at (B) for a 19:2 ratio.

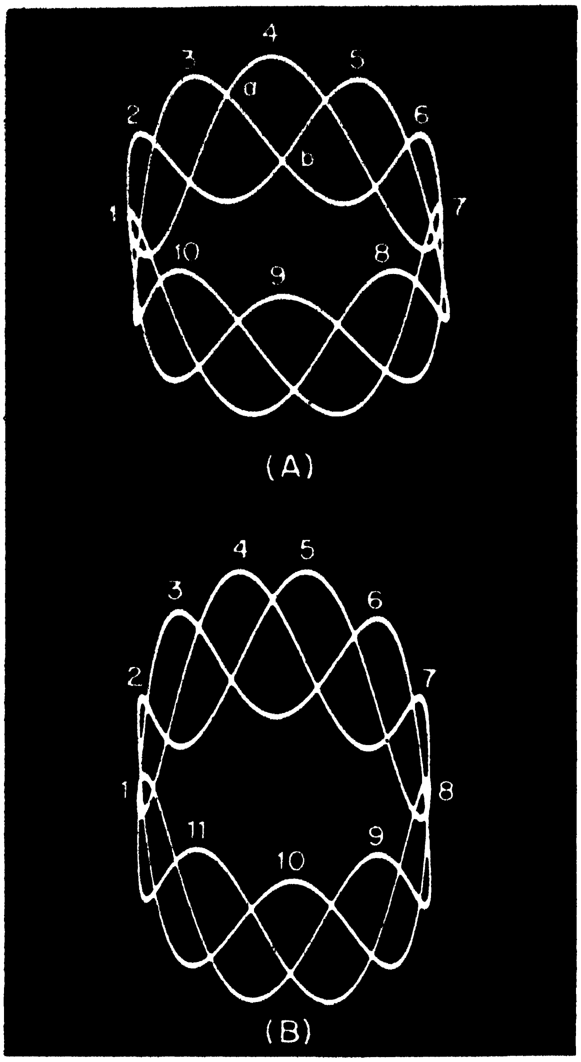

Fig. 12-93.—Frequency ratios of 10:3 and 11:3 are indicated by (A) and (B) respectively.

In Fig. 12-93 appear two other patterns produced by the method indicated in Fig. 12-90. The peaks of each pattern are marked as before, so that the numerator of the ratio for the pattern at (A) is 10 and for that at (B) is 11. The denominator of the ratio for each pattern is 3, because for each image there are two intersecting lines to which we add 1. In order to show two such intersecting points, we have arbitrarily marked one part of the pattern of Fig. 12-93A with letters *a*

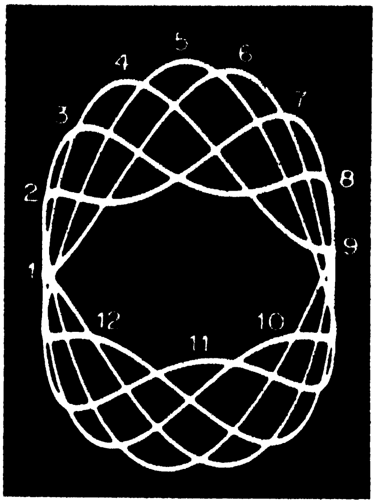

Fig. 12-94.—A frequency ratio of 12:5 is indicated here.

and *b*. In other words, that part of the pattern leading from peak number 3 and going in the negative direction (i.e., with the negative slope) has only two intersecting lines crossing its path. Consequently, Fig. 12-93A indicates a 10:3 ratio, while Fig. 12-93B shows an 11:3 ratio.

A higher ratio pattern is indicated in Fig. 12-94. By carefully counting the peaks (which we have numbered), 12 of them can be seen. The number of intersections for any one line is four, which means that the denominator of the frequency ratio is equal to five, and hence the frequency ratio represented by the pattern is 12:5. In all of the patterns shown in Figs. 12-91 through 12-94, the higher-frequency (modulating) signal was applied to the vertical-deflection plates. This is easily noticeable by the sine waves of the pattern being deflected in the vertical direction. If this higher-frequency signal were applied to the horizontal-deflection plates, the patterns shown here would be shifted in position by 90°.

A very simple equation can be formed for determining the frequency ratios from such patterns as obtained by the circuit arrangement of Fig. 12-90. If we let *P* equal the number of peaks around the *top*, or *circumference*, of the pattern (visualizing the image as rotating around a cylinder) and *n* the number of intersecting lines (obtained as outlined before), then the ratio can be expressed as follows

$$\text{Frequency ratio} = \frac{P}{n+1}\ . \qquad (12\text{-}5)$$

At all times, the numerator of the equation will be greater than the denominator.

Another form of circuit[7] utilizing the same phase-splitting network as before is shown in Fig. 12-95A. This circuit produces a pattern such as that in Fig. 12-95B. The voltage is effectively applied across both sets of plates in quadrature (i.e., 90° out of phase) by means of the phase-splitting circuit. As a result of this circuit, when the reactance of *C* is equal to the resistance of *R*, the beam spot traces a circle upon the screen. The other voltage, the unknown, is connected in series with the accelerating anode (anode no. 2). The operation of the system is predicated upon the fact that as a result of the varying voltage in series with the fixed accelerating voltage, the anode potential will fluctuate through a maximum and minimum, in accordance with the frequency of the varying voltage. At the same time, the voltage applied in quadrature to the two input terminals, and hence to the two sets of deflection plates, will cause the spot to describe a circle. The varying voltage will cause the anode potential to fluctuate a number of times (depending upon the frequency of the varying voltage) during each circulation of the spot. This spot movement is different from that accomplished by the circuit of Fig. 12-90. For the circuit of Fig. 12-95, the spot, as it travels around the circle, is effectively displaced in a radial manner relative to the circle. The result is that the final pattern will be an image which indicates the frequency ratio between the varying voltage in series with the anode and the frequency of the voltage which is causing the circular motion of the spot. If the ratio between the two is integral, the image will be stationary; if the ratio is fractional, the image will revolve.

The frequency ratio for such patterns is determined in a similar fashion to that just discussed. The number of peaks on the outside (or inside) of the pattern is the numerator of the ratio and the number of intersections plus one is the denominator of the ratio. Thus the pattern of Fig. 12-95B represents a 19:2 ratio. This is tantamount to stating that the ratio is equal to the number of whole sine waves superimposed around the circle for any one revolution divided by the number of revo-

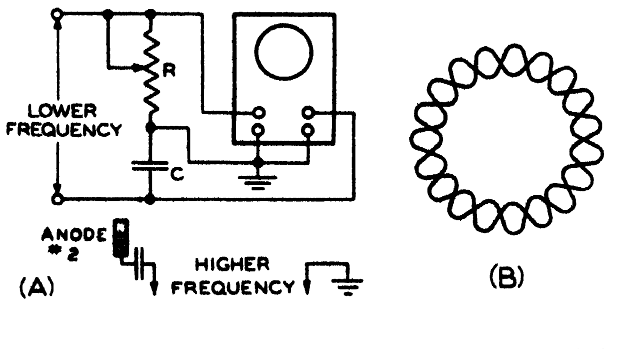

Fig. 12-95.—In this method of frequency comparison (A), the higher-frequency signal is fed to the second anode of the cathode-ray tube; the resultant pattern is for a 19:2 ratio.

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[7] Kipping, N. Y., "Waveform examination with the cathode-ray oscillograph," *Elec. Commun.*, vol. 3, pp. 69-75, July, 1924.

lutions the spot makes around the circle (i.e., the number of times the same complete sine wave is repeated).

This arrangement for frequency comparison is seldom used. It has several disadvantages when viewed from the practical angle. First is the danger entailed by working around the high-anode voltage found in present-day cathode-ray-tube equipment. The second is that it is quite a problem to break into the finished cathode-ray oscilloscope in order to make proper contact with the circuit. The third is the defocusing effect of this varying potential, and the resultant blurring of the image.

Roulette Method[8-12]

There is still another method by which frequency comparison can be made. The resulting patterns from this method are shaped in the form of a roulette, and hence it is often called the *roulette method*. In the circuit arrangement to be discussed, *two* separate phase-splitting networks are used.

In the frequency-comparison circuits previously described, where a phase-splitting network was employed, no mention was made as to the type of arrangement of the deflection plates. In other words, one plate from each pair can be internally connected together (and then grounded), or they can all have separate connections. In the method to be described in this section, both types of deflection-plate arrangements can be used, but where one plate from each pair is internally connected together, additional circuit components are needed to produce the same effects as for the type with separate external connections. We will first study the circuit arrangement where each deflection plate has external connections.

The circuit arrangement appears in Fig. 12-96. The input signals can be fed directly to the cathode-ray tube as shown (the signals must be strong enough to produce readable deflections), or they can be fed through the amplifier sections of the oscilloscope. In the latter case, the cathode-ray tube must have the plates connected in a balanced deflection system. Note the two *R-C* phase-shifting networks. The resistors and capacitors are assumed to be pure, that is, the resistors are assumed to contain negligible inductance and the capacitors negligible leakage resistance. The two input signals are introduced as shown.

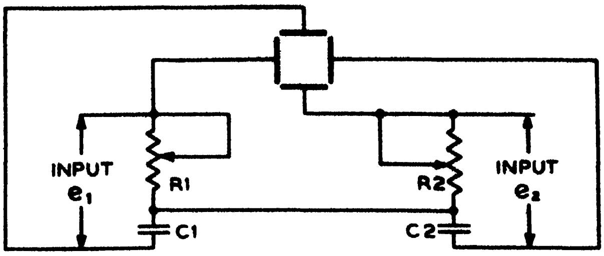

Fig. 12-96.—Schematic circuit arrangement for the roulette method of frequency comparisons.

The resistors are made variable so that after proper adjustment, each signal when considered acting alone will produce a circular trace on the screen of the cathode-ray tube. If signal voltage e_1 is the only active input signal, the voltage drop across *R1* is coupled to the right horizontal-deflection plate through *C2* and the voltage drop across *C1* is coupled to the lower vertical-deflection plate through *R2*. A similar action will occur when signal voltage e_2 is considered acting alone. The strength of the input signals determines the radius of the individual circles produced. For proper operation of this system, the circle produced by the lower-frequency signal should be larger than, or just equal to, that of the higher-frequency circle. This is necessary because the simultaneous combination of the two circular forces will be such that the spot will traverse a path which may be visualized as a point on the small circle rotating about the circumference of the larger one. This combined action is such that there will be deviations or breakups in the regular path of the lower-frequency circle (the deviations are the same all along the path) which will be a ready means of determining the frequency ratio.

For the circuit illustrated in Fig. 12-96, there will be loops, points, or cusps that break outward, but if the resistor and capacitor of *either* phase-splitting network are interchanged, the loops will be pointed inward. In this latter case, there will be fewer cusps or loops, assuming the frequency ratio is the same for both cases. For ratios where the higher frequency is an integral multiple of the lower frequency (i.e., where the ratio in its simplest form is an integer) it is easy to determine the ratio, whether the loops are pointed inward

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[8] Dye, D. W., "Improved cathode-ray tube method for the harmonic comparison of frequencies," *Proc. Phys. Soc.*, vol. 37, p. 158, April 15, 1925.

[9] Rangachari, T. S., "The harmonic comparison of radio frequencies by the cathode-ray oscillograph," *Exper. Wireless and Wireless Eng.*, vol. 5, p. 264, May, 1928.

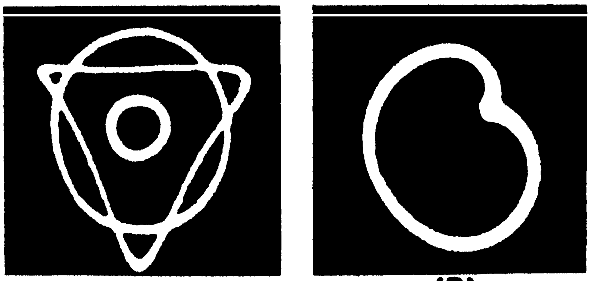
[10] Rangachari, T. S., "The super-position of circular motions," *Experimental Wireless and Wireless Eng.*, vol. 6, p. 184, April, 1929.

[11] Kurrelmeyer, B., "A simple circuit for oscillographic frequency comparison," *Rev. Sci. Instr.*, vol. 7, p. 200, May, 1936.

[12] Reich, H. J., "Circuits for oscillographic frequency comparison," *Rev. Sci. Instr.*, vol. 8, p. 348, September, 1937.

or outward. When the ratio is not an integral value, it is best to have the loops pointed outward, because the other type of pattern becomes more difficult to interpret. Let us study some typical oscillograms before we discuss how the frequency relationship is calculated.

The photo of Fig. 12-97A is a triple exposure and is based upon the circuit of Fig. 12-96 as is. The small circle is that produced with signal e_2, the higher-frequency voltage, acting alone, and the large circle is produced by the lower-frequency voltage e_1 acting alone. In other words, the amplitude of e_1 is greater than that of e_2. The other part of the photo is the resultant pattern caused by the combined action of the two input signals, e_1 and e_2. This resultant pattern is for a frequency ratio of 2:1. If the resistor and capacitor of either phase-splitting network of Fig. 12-96 are interchanged, the resulting pattern will have its peaks pointed inward as seen in Fig. 12-97B.


(A) **(B)**

Courtesy Rev. Sci. Instr.

Fig. 12-97.—The resultant pattern of the triple exposure photograph of (A) and the pattern of (B) are both 2:1 frequency ratios by the roulette method.

If the amplitude of the higher frequency is increased and approaches that of the lower frequency, the peaks or points of the resultant patterns of Fig. 12-97 will open up and loops will be formed. This is best illustrated by the 2:1 ratio of Fig. 12-98. Compare this photo with the resultant of Fig. 12-97A. Both oscillograms can be used equally well for frequency comparison. If the amplitude of each input signal were the same, the loops would open up more and the spot would cross the *center* of the pattern three times.

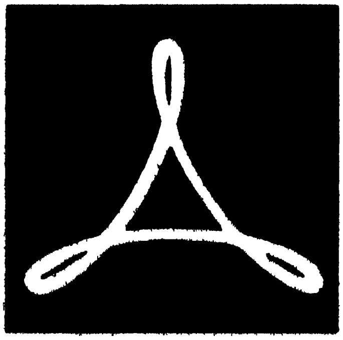


Fig. 12-98.—A roulette with a 2:1 frequency ratio; the loops indicate that the amplitude of the higher-frequency signal approaches that of the lower-frequency signal.

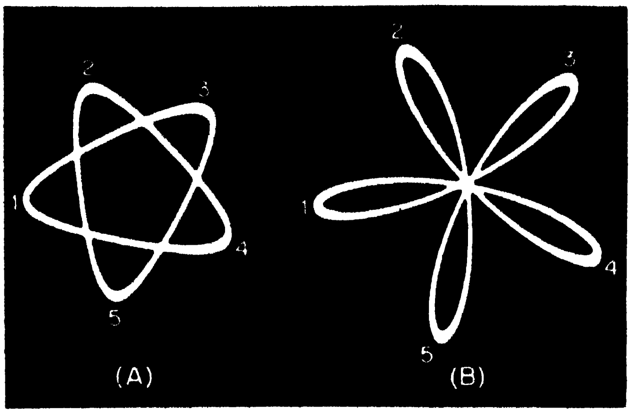


Fig. 12-99.—Both patterns indicate roulette frequency ratios of 3:2.

The two patterns of Fig. 12-99 each indicate a frequency ratio of 3:2. The pattern at (A) resulted when the amplitude of the higher-frequency signal was lower than that of the lower-frequency signal. When the amplitudes of both signals were made approximately equal to each other, the peak opened up into loops, and the pattern at (B) resulted. Note from this latter photo that the loops all meet at the center of the pattern. The spot crosses the center of the pattern five times.

From the pictures we have seen thus far, we can formulate a method whereby we can determine the frequency ratio from the pattern. If in tracing out the pattern, the spot goes from one loop or peak directly to the next (whether the peaks are pointed inward or outward), then the denominator of the ratio will always be unity and the resultant ratio will be an integral value. The patterns of Figs. 12-97 and 12-98 illustrate such a condition. If the spot does not go from one peak to the next, but skips some, the denominator of the ratio is different from unity (see Fig. 12-99). If N_H/N_L is the ratio of the higher to lower frequency in its simplest form, then irrespective of its magnitude, as long as the pattern has the cusps or loops pointed *outward*, the total number of loops, call it N, directly indicates the sum of $N_H + N_L$. Thus

$$N = N_H + N_L. \tag{12-6}$$

Solving equation (12-6) for N_H we find

$$N_H = N - N_L. \tag{12-7}$$

Consequently, the frequency ratio N_H/N_L can now be expressed as follows

$$\text{Frequency ratio} = \frac{N - N_L}{N_L}. \tag{12-8}$$

The value of N_L for those cases where the spot traces the pattern by going from one cusp or loop directly to the next, as we already mentioned, is equal to unity. From Fig. 12-99, which indicates 3:2 ratios, you will note that the spot in going from any cusp or loop to the next can be considered as skipping *either* one or two loops. Thus from loop *1* (the loops are marked for the sake of illustration) in either pattern, we find that the spot may be traced toward loop *3* or *4*. In going toward loop *3*, the spot can be considered as skipping either loop *2* on one hand or loops *4* and *5* on the other hand. In going from loop *1* toward loop *4*, however, we either skip loop *5* or the two loops numbers *2* and *3*. Regardless of the pattern or the direction in which we follow a loop, two groups of loops will be skipped. Also, regardless of which cusp or loop we start with, the same situation will always hold. If the denominator of the ratio is increased, then the smaller number of skipped loops will also increase. The arrangement of the pattern produced by the circuit of Fig. 12-96 is such that the denominator or the N_L of the ratio can be determined from any pattern (in which the loops are pointed *outward* by adding *one* to the *smaller* number of loops skipped. Thus for Figs. 12-97 and 12-98, there are no loops skipped, so N_L is 1, while for Fig. 12-99, the smaller number of loops skipped is 1, and hence N_L for this case is 2.

Getting back to equation (12-8) we can easily determine the frequency ratios from the pattern. For Figs. 12-97A and 12-98, N (the total number of loops) equals 3 and N_L equals 1, so that the ratio becomes

$$\frac{N-N_L}{N_L}=\frac{3-1}{1}=\frac{2}{1}$$

and for Fig. 12-99, N equals 5 and N_L equals 2, hence the ratio is

$$\frac{N-N_L}{N_L}=\frac{5-2}{2}=\frac{3}{2}\,.$$

If the cusps or loops are pointed inward, then the minus sign of equation (12-8) is changed to a plus sign. Thus for Fig. 12-97B, N equals 1, N_L also equals 1, and the ratio becomes

$$\frac{N+N_L}{N_L}=\frac{1+1}{1}=\frac{2}{1}\,.$$

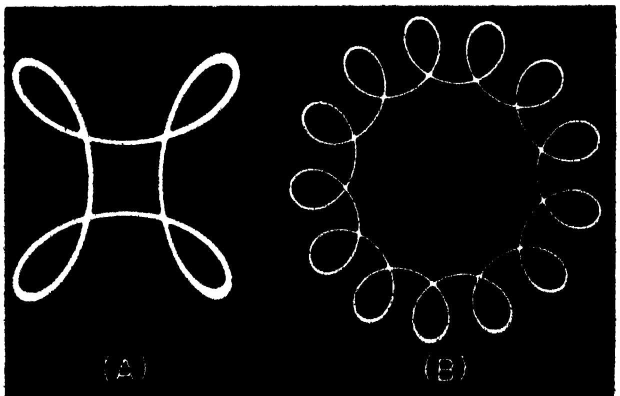


Fig. 12-100.—Integral ratio roulette patterns; (A) indicates a 3:1 frequency ratio and (B) a 12:1 ratio.

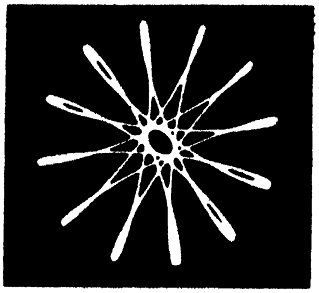


Fig. 12-101.—A roulette pattern for a 7:5 frequency ratio.

For frequency ratios other than integral values, it is not advisable to use the arrangement where the cusps point inward, because the pattern becomes quite intricate, and it becomes difficult to interpret N and N_L.

Let us study some other patterns produced by the circuit of Fig. 12-96, where the peaks or cusps are pointed outward. The amplitudes of the input signals of these forthcoming patterns are such that all the peaks are opened up. The two patterns of Fig. 12-100 indicate integral ratios, where N_L is equal to unity. For the roulette pattern at (A) 4 loops are counted, so the frequency ratio becomes 3:1, and in pattern (B), 13 loops are counted, so the ratio is 12:1. For the pattern of Fig. 12-101, N equals 12. Now if we choose the extreme end of any one loop as a fixed point and follow the path of the trace in either direction to the extreme end of another loop, which we take as our second fixed point, we shall find that we have skipped two groups of loops between these two fixed points. One group comprises 6 loops, the other 4 loops. Hence N_L which is 1 more than the smaller number of loops skipped, is $1+4=5$, and the ratio becomes 7:5. For Fig. 12-102, the ratio is 10:7. This is easily seen by counting a total of 17 loops and 6 skipped loops (the smaller number of skipped loops). There are 23 total loops in the pattern of Fig. 12-103, and if careful observation is made, we will find that the *smaller* number of loops skipped is 10, which means that the ratio is 12:11.

A disadvantage of this system of frequency comparison is that we cannot determine directly from the pattern which signal input is the higher frequency. However, if only the higher frequency is increased, the greater number of loops skipped will increase but not the smaller, and if the lower frequency is increased, the

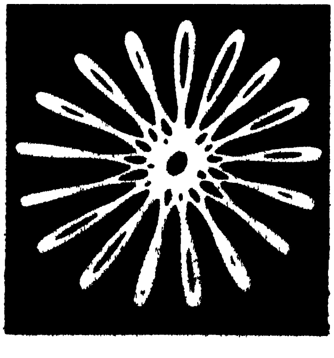


Fig. 12-102.—A frequency ratio of 12:7 is indicated by this roulette pattern.

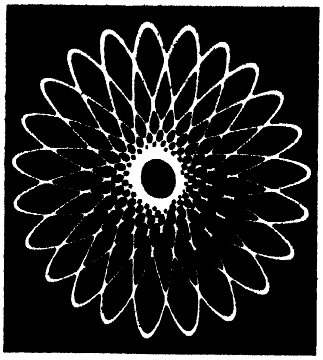


Fig. 12-103.—A roulette pattern indicating a 12:11 frequency ratio.

smaller number of loops skipped will increase and not the larger. Thus if we increase the frequency of either the known or unknown signal to the next stationary pattern and note the change in the number of loops skipped, we have a ready means of knowing which input signal is of higher frequency.

The analysis of roulette patterns thus far is based upon the original work done by the authors of the reference articles.[8–12] In the process of working with such patterns and obtaining the patterns indicated in this section, this author has found a supplementary method of determining the frequency ratio from the patterns. This method is also based upon the fact that the total number of loops or cusps of the resultant pattern is equal to the sum of the numerator and denominator of the ratio (i.e., it is based upon $N = N_H + N_L$).

When the ratio is an integral value, no loops are skipped, and the calculation of the ratio is easily done from the pattern. When the ratio is a fraction in its simplest form, then some loops will be skipped when the spot traces out the pattern in going from one loop to another. We have shown that *two* groups of loops are skipped, one group containing more loops than the other. It is upon both of these groups that this supplementary method of determining the frequency ratio is based.

If we let N_G equal the group which contains the greater number of loops skipped, and N_S the group which contains the smaller number of loops skipped, we can write

$$N = N_G + N_S + 2. \qquad (12\text{-}9)$$

This is so because in counting N_G and N_S we omit the two loops which we use as our fixed or starting points. If we take equation (12-9) and substitute it for the N in equation (12-8), and if we let $N_L = N_S + 1$, we find that

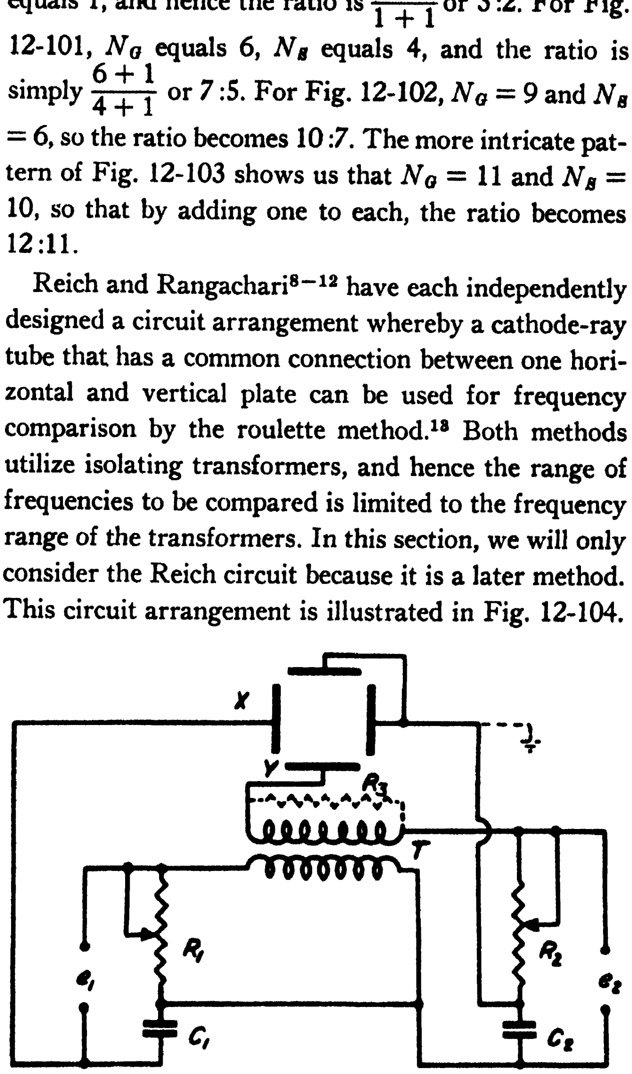
$$\text{Frequency ratio} = \frac{(N_G + N_S + 2) - (N_S + 1)}{N_S + 1}$$

which simplifies to (12-10)

$$\text{Frequency ratio} = \frac{N_G + 1}{N_S + 1}\,. \qquad (12\text{-}11)$$

Equation (12-11) simply states that the frequency ratio of roulette patterns, ***with cusps or loops pointed outward,*** is equal to 1 plus the greater number of loops skipped, divided by 1 plus the smaller number of loops skipped. Thus for Fig. 12-99, N_G equals 2, and N_S equals 1, and hence the ratio is $\frac{2+1}{1+1}$ or 3:2. For Fig. 12-101, N_G equals 6, N_S equals 4, and the ratio is simply $\frac{6+1}{4+1}$ or 7:5. For Fig. 12-102, $N_G = 9$ and $N_S = 6$, so the ratio becomes 10:7. The more intricate pattern of Fig. 12-103 shows us that $N_G = 11$ and $N_S = 10$, so that by adding one to each, the ratio becomes 12:11.

Reich and Rangachari[8–12] have each independently designed a circuit arrangement whereby a cathode-ray tube that has a common connection between one horizontal and vertical plate can be used for frequency comparison by the roulette method.[18] Both methods utilize isolating transformers, and hence the range of frequencies to be compared is limited to the frequency range of the transformers. In this section, we will only consider the Reich circuit because it is a later method. This circuit arrangement is illustrated in Fig. 12-104.



Courtesy Rev. Sci Instr.

Fig. 12-104.—Schematic arrangement whereby roulette patterns can be formed on a cathode-ray tube which has a common connection between one deflection plate of each pair.

As before, both phase-splitting networks are adjusted so that each input signal, when acting alone, produces a circular pattern. The following is an excerpt from the article by Reich,[12] with reference to the circuit of Fig. 12-104.

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[18] If such tubes were used without any special circuit components, then, according to Reich, "the internal connection between the two deflection plates either shortcircuits one of the input voltages or connects each voltage directly to one pair of plates resulting in the production of Lissajous figures, instead of roulettes."

"The impedance of the isolating transformer T should be high in comparison with the resistance R_1. For frequencies in the audible range, excellent results are obtained with an ordinary interstage coupling audio transformer and a 5,000-ohm variable resistance. At frequencies above 5,000 cycles, the distributed capacitance of the transformer tends to prevent the formation of a circular pattern for the voltage e_1. For this reason, e_1 should be the lower frequency. Resistance or capacitance between the Y plates or leads produces a phase shift which makes it difficult to obtain a circular pattern for the voltage e_2 at very high frequencies. No difficulty is encountered, however, even above 10,000 cycles, if the plates are not shunted by a leak and if the connection from the ungrounded plate to the transformer is short. A high resistance R_2 across the secondary is sometimes of help at high frequencies. If the primary and secondary of the transformer of Fig. 12-104 are wound in the same direction, the cusps or loops of the pattern will point inward. The pattern may be transformed into one in which the cusps point outward by transposing the primary or secondary connections or by interchanging the position of the capacitor of either R-C phase-shifting network."

It is evident that considerable progress has been made in audio transformer development, since this early (1937) article has appeared; very little difficulty due to distributed capacitance should be experienced with present-day transformers throughout the audio range.

Oscillator Calibration

Two methods of calibration are possible, using the cathode-ray tube as a visual indicator. One is to adjust the known and the unknown frequencies of the oscillators so as to produce a 1:1 pattern upon the screen; for this adjustment, the unknown oscillator frequency is equal to the known frequency of the standard oscillator. The other method is to keep the standard oscillator fixed at one frequency, say the lowest base frequency, and to vary the unknown oscillator through a range which will produce patterns from a 1:1 ratio on up (according to how high the frequency of the calibration is desired), calibrating each adjustment of the unknown, in accordance with the frequency ratio indicated upon the screen. (The method of frequency comparison to be used, whether by Lissajous figures, intensity modulation, roulette method, etc., depends upon the range of calibration and the frequency ratios to be dealt with.)

The former of these two calibrating methods is to be preferred. First, because it enables more accurate adjustment at each frequency, since the frequency ratio is 1:1. Second, it minimizes the error to that inherent in the standard at the particular frequency. If the second method is used and the base-frequency adjustment is "off," the discrepancy is carried on through each of the other calibrations based upon the original standard frequency. However, the second method is faster, since only one oscillator is adjusted for a number of frequency calibrations. It is more economical and perhaps more common because of the minimum of expense involved to secure a single-frequency standard. If the first method is used, each oscillator is advanced step by step. The closer the calibration points, the greater is the number of tuning changes required.

The degree of accuracy obtained is determined by two factors. The first is the intrinsic accuracy of calibration of the standard oscillator for any one setting. The second is the care with which the unknown oscillator is varied in order to stop the image. Absolute accuracy of calibration is achieved when the image is stopped. However, it is possible that either one of the two oscillators used will drift, thus preventing the stopping of the image. At any rate, a very slow drift of the pattern is permissible and often unavoidable. The point of calibration should be set as that for which the image has stopped or is moving at the slowest obtainable speed.

Very little need be said about calculations for 1:1 ratio patterns. When these appear and are stopped, the unknown oscillator frequency is equal to the standard frequency. If the calibration of the unknown oscillator ranges is carried out completely in this manner, there is no need for calculation of any sort. However, if the frequency calibration is carried out with ratios other than unity, a simple calculation is required. It is best in most cases to make the standard frequency the base frequency and lower than any unknown frequency. This means that the pattern will indicate ratios, wherein the unknown frequency is always the greater. Thus if a pattern indicates a 4:1 ratio, the lower frequency is the standard and the higher frequency is the unknown. If the base or standard frequency for this pattern is, say 100 cps, the unknown frequency for the setting which develops the 4:1 ratio pattern is 100 × 4/1 or 400 cps. In other words, the base frequency is multiplied by the ratio obtained on the oscilloscope. Thus, as another example, if the base frequency (standard) is 1,200 cps and is the lower of the two frequencies which produced the pattern, and the frequency ratio indicated by the pattern is 4:3, the frequency of the unknown for that setting is 1,200 × 4/3 or 1,600 cps.

For a base frequency of 100 cps, the following table lists the frequencies of the unknown corresponding to various ratios, assuming that the base frequency is always lower or equal to the unknown frequency.

Ratio	*Frequency of Unknown Oscillator (cps)*	*Ratio*	*Frequency of Unknown Oscillator (cps)*
1/1	100	9/4	225
5/4	125	7/3	233
4/3	133	5/2	250
3/2	150	8/3	267
5/3	167	11/4	275
7/4	175	3/1	300
2/1	200		

The corresponding patterns for these thirteen ratios, produced for various base frequencies, will enable calibration in extremely small steps. Thus, suppose that the thirteen ratios listed have been established. By increasing the base frequency to 300 cps, as the second adjustment, the ratios listed will provide calibration points for 300 cps, 375 cps, 400 cps, 450 cps, 500 cps, 525 cps, 600 cps, 675 cps, 700 cps, 750 cps, 800 cps, 825 cps, and 900 cps. By moving the base frequency up to 900 cps and producing again the same ratio patterns, calibration up to 2,700 cps is obtained. Similarly, by advancing the base frequency up to 2,700 cps and producing the same ratio patterns by adjusting the unknown oscillator frequency, calibration up to 8,100 cps is effected. This process can be carried on indefinitely, until all ranges of the oscillator are calibrated against the standard.

CHAPTER 13

NONLINEAR TIME BASES

In Chapter 9, we discussed linear time bases separately because linear sweep circuits are the type most commonly used in oscilloscopes. There are, however, numerous other types of time bases used in special kinds of oscilloscopes and other equipment where cathode-ray tubes are employed. To distinguish these time bases from the types that produce a sawtooth waveform (or modification thereof), we will call them *nonlinear time bases.* The most common types of nonlinear time bases are often classified as *circular* (or *polar*), *spiral, elliptical,* and *radial* sweeps. We will study these types of time bases in this chapter.

The term polar sweep, or polar co-ordinate sweep, is often misrepresented and misunderstood by readers. When a circular, elliptical, or spiral time base is said to represent a polar sweep, it means that the trace on the screen produced by the time base can be considered as being shown on a system of polar co-ordinates. Signals can be displayed on such traces in many different ways. Generally speaking, a polar co-ordinate representation is one in which the radius (or radius vector, as it is sometimes called) varies in both magnitude and direction to trace out the curve in question. On the basis of this definition, any signal displayed along a circular or spiral time base is a polar co-ordinate representation. However, very seldom can a signal portrayed by such time bases be considered as a *plot* on polar co-ordinates in the true sense of the word, because this portrayal does not represent the mathematical relation between the magnitude and direction of the radius vector the same way as on polar co-ordinate graph paper. In other words, a plot of a sine wave on polar co-ordinate graph paper will be a completely different pattern than a sine wave spread out by these time bases.

Circular and Elliptical Sweeps

There are a variety of methods for producing circular and elliptical sweeps, some of which we have already studied in Chapter 12 when considering diverse means of frequency measurements. One of the chief reasons for employing a circular (or elliptical) time base is to increase the over-all length of the trace. Thus, if the circular trace has a diameter of four inches, the circumference of the circle will be 4π, or 12.56 inches. From this we see that a circular trace of four-inch diameter will have an effective length π times (3.14 times) that of a linear (straight-line) trace of four inches. Consequently, we see that more information can be displayed on a circular time base than one produced by a sawtooth sweep. Another advantage of such time bases over sawtooth sweeps is that no time is lost on retraces. Circular time bases, in addition to being used for frequency comparisons, also find a wide variety of uses in radar.

In Chapter 12, we have shown that by applying a sine wave to each set of deflection plates, the resulting pattern will be a circle when the two sine waves are 90° (or 270°) out of phase and the amplitudes of the two input signals are equal (see Figs. 12-15, 12-16, and 12-17); a symmetrical ellipse will result when the amplitudes are not equal (see Figs. 12-18 and 12-19). We have also shown that such traces can be made by a single sine-wave source, by applying this signal to an *R-C* phase-splitting network and feeding the output of this network to the input terminals of the oscilloscope (see Fig. 12-87).

In Chapter 12, we have seen how circular traces can be used for frequency measurements. Circular and elliptical traces are also used just as a time base to portray any type of signal, be it of sinusoidal character, a complex wave, or a pulse. The (A) and (B) patterns of Fig. 12-91 in Chapter 12, although they are used to portray frequency ratios of 10:1 and 12:1, respectively, can also be thought of as illustrating a sine wave spread out by a circular trace. If the circular trace is made elliptical, and the signal input to the horizontal-deflection plates is increased so that part of the pattern is removed from the screen, the image seen will be that of two *separate* waves. By proper manipulation, certain parts of the input signal can be observed easily in comparison with other parts. It should be remembered that for a signal to be properly presented along a circular or elliptical time base, its frequency should be an integral multiple of the signal frequency causing the circular trace.

Injecting the signal to be observed to any one pair of deflection plates (see Fig. 12-90 in Chapter 12) to produce the patterns mentioned above, the deflection of the signal by the circular time base will *not* always be at right angles to this circular trace, but may actually be tangent. This is especially obvious at the extreme right and left sides of the patterns of Fig. 12-91. In other words, the reproduction of a signal by a circular time base by the method just described will make it difficult to interpret certain parts of the reproduced picture.

Radial Deflection of the Circular Trace

If the signal to be observed can cause a deflection of the spot that *will always* be at right angles to the circular time base, that is, where the deflection will always be radial, we will have a superior arrangement of using this circular time base. There are various methods whereby such deflection is possible. One we discussed briefly in Chapter 12 (see Fig. 12-95) where the circular trace was produced by a simple *R-C* phase-splitting network and where the signal to be observed was fed to the second or final anode of the cathode-ray tube, and hence modulated the circular trace in a radial manner. We also outlined the basic disadvantages of such a system and indicated that although it produced the required results, they were not obtained easily.

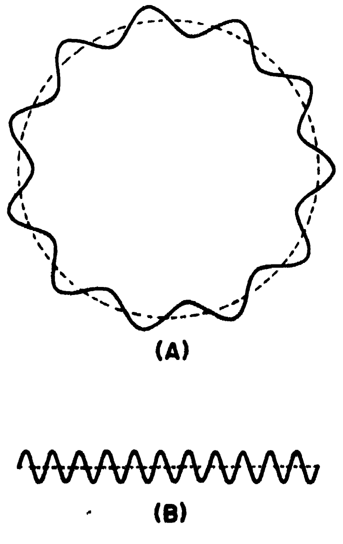

Fig. 13-1. — A sine wave is shown displayed by a circular trace with radial deflection in (A), and in (B) by a linear trace whose length is equal to the diameter of the circular trace.

Whatever the case may be, if only one trace of a signal is to be observed and if the signal is periodic, then its frequency must be an integral multiple of that causing the circular time base. The drawing of Fig. 13-1A illustrates a sine wave spread out by a circular trace in which the deflection along the trace is radial. The method of producing such a trace is, at the moment, unimportant. What is important is the fact that the complete sine wave is visible and no portion is cramped. The dashed circle of this pattern indicates the position of the circular trace before application of the modulating signal. You will also notice that 11 cycles of the sine wave are spread out along the circular baseline. This means that the frequency of the signal being observed is eleven times as great as that producing the time base.

In part (B) of Fig. 13-1 11 cycles of the same sine wave are reproduced by a linear time base, with the length of the time baseline being equal to the diameter of the circular trace at part (A). By comparing these two drawings, you can readily see how a circular time base spreads out the waveform for easier observation. For the same time separation between cycles, a circular trace can occupy π (or 3.14) times as many cycles as a linear trace whose length is the same as the diameter of the circle.

There are two chief methods whereby radial deflection of a circular trace can be effected; one involves the use of additional electrode elements inside the cathode-ray tube which are made to function in conjunction with the deflection plates, and the other employs an amplifier arrangement in conjunction with the deflection plates. In the following sections, we will study a number of these different arrangements for producing a circular trace and modulating it in a radial manner.

Amplifier Arrangement for Radial Deflection

One of the first methods devised for causing radial deflection of a circular time base involved the use of a balanced amplifier system. This method dates back to 1932 and was initially developed in England.[1] In order to utilize the circuit (Fig. 13-2) to the best advantage, a cathode-ray tube in which each deflection plate has a separate external connection is employed. Let us examine this circuit and see how it produces a circular time base as well as radial deflection of trace, when the signal to be observed is applied.

The sinusoidal signal used to produce the circular time base by means of two 90° out-of-phase signals is applied across a resistance-capacitance phase shifter, *R1-C1* in Fig. 13-2. The reactance of capacitor *C1* is equal to the resistance of *R1* at the frequency of the applied signal, so that the voltages developed across

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[1] The Staff of The Radio Research Station, Slough, "A circular time-base giving radial deflections, for use with the cathode-ray oscillograph," *Jour. IEE* (London), vol. 71, p. 82, June, 1932.

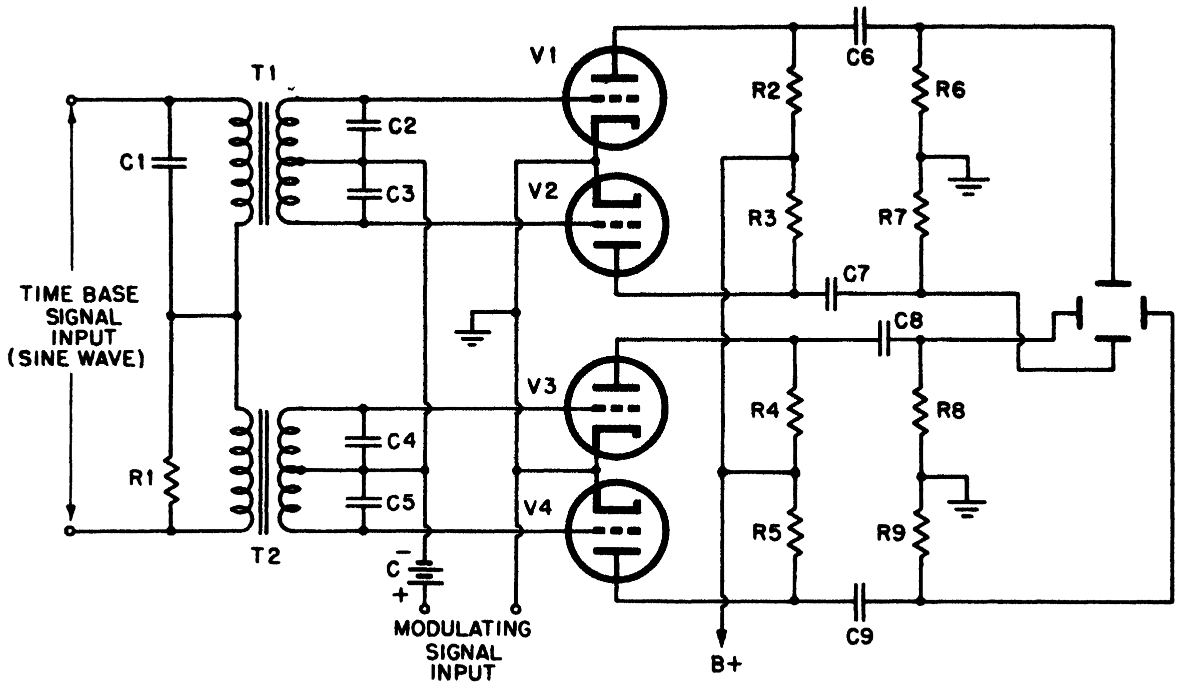

Fig. 13-2.—Balanced amplifier circuit for producing radial deflection for a circular time base.

both components are equal, resulting in a circular and not an elliptical time base. Thus, we have the two equal 90° out-of-phase voltages to produce the circular trace, but you will notice that they are applied to the deflection plates through two pairs of amplifying tubes connected for push-pull operation. The tubes are employed primarily as a means of introducing the modulating signal (the signal to be observed) in such a manner that it will radially deflect the circular time base. Before we discuss how the modulating signal is introduced, let us see how the amplifier circuit keeps the signals applied to the deflection plates 90° out of phase.

Let us first consider the circuit consisting of push-pull amplifiers *V1* and *V2*. The voltage appearing at the top of the secondary winding of *T1* is 180° out of phase with that appearing at the bottom, since equal and opposite voltages are induced into the secondary with respect to the center tap; hence, equal, but 180° out-of-phase voltages appear at the grids of *V1* and *V2*. Due to the inherent phase-inversion quality of an amplifier, the signals appearing at the plates of these two tubes are 180° out of phase with the signals at their respective grids. This means that the signal between the plate circuits of *V1* and *V2* have the same polarity as that across the primary of *T1*.

Resistors *R2* and *R3* serve as the plate loads for tube *V1* and *V2*, respectively. The output signal from each one of these tubes appears across its load resistor and is coupled to the vertical-deflection plates as seen in Fig. 13-2. Capacitor *C6* serves to couple the signal from *V1* to one vertical plate, and capacitor *C7* couples the signal from tube *V2* to the other deflection plate. Thus we see that the signal appearing across capacitor *C1* of the input phase-splitting network finally appears amplified in magnitude across the vertical-deflection plates.

Since the bottom half of the circuit, consisting of push-pull amplifiers *V3* and *V4*, is identical to the upper half, we can readily understand that the signal appearing across resistor *R1* also appears across the horizontal-deflection plates amplified by the same amount. Consequently, we find that the two 90° out-of-phase signals appearing across the *R1-C1* phase-shift network also appear across the deflection plates in the same relative phase, since possibly phase shifts introduced by capacitors, etc., will be identical for the upper and lower halves of the circuit, and therefore, will not affect the 90° *relative* phase between the vertical and horizontal plates.

The circuit must be perfectly balanced if a true circular trace is to be formed. In other words, center-tap transformers *T1* and *T2* must be exactly the same so that the signal inputs to the four tubes are the same in magnitude and of the proper phase; the tubes themselves must all be the same and accurately matched. The plate and grid voltages on the tubes must be the same; the plate-load resistors *R2, R3, R4,* and *R5* must all be equal, as well as coupling capacitors *C6, C7, C8,* and *C9.* Resistors *R6, R7, R8,* and *R9* are all equal and help match the deflection plates to the amplifier circuit. Of course, with the circuit balanced, the deflection sensitivities of each pair of deflection plates must also be the same to produce a circular trace. However, since cathode-ray tubes exhibit slight differences in deflection sensitivity between the plates due to their geometrical construction, these must be taken care of by proper circuit adjustments.

The grids of all the tubes have the same bias applied by being connected to the *C* battery, as indicated in Fig. 13-2. The signal to be examined, that which will radially modulate the circular trace, is applied to the grids in series with the bias voltage. The tubes are biased to the point where they will be operating over a nonlinear portion of their i_b-e_g transfer characteristic. This is a necessary condition for the proper function of this circuit. As in most modulation systems, for the modulating signal to show up in the output in conjunction with the carrier, or time-base signal in this case, sum and difference frequencies must exist in the output circuit. Such sum and difference frequencies are brought about only by operating the tubes over a nonlinear portion of the transfer characteristic. If the linear portion of the transfer characteristic were used, then the modulating signal, which is applied in the same phase to the grids of each tube, would cancel out in the output load circuit, because equal and opposite currents would flow through the load resistors.

The signal to be observed should be of a higher frequency than the time base. Capacitors *C2, C3, C4,* and *C5,* which are all of the same value, serve as a low-impedance path to this input modulating signal. Since the signal to be observed is applied in the same phase to each grid, it by itself will not produce any deflection of the spot, but when acting in conjunction with the time-base signal, it will modulate this latter signal in the four tubes. Hence the signal to be observed in effect controls the diameter of the circle, thereby causing radial deflection.

The radial deflection of the circular trace is proportional to the instantaneous voltage of the applied modulating signal. Assuming the signal to be observed is a sine wave, then during its positive half-cycles, the instantaneous bias on the grid will decrease (i.e., the bias will become less negative), thereby causing the instantaneous radius of the circle to *increase* from its steady-state value, in proportion to the amplitude of the positive half-cycle of the sine wave. During the negative half-cycle, the instantaneous bias will increase proportionately, causing the radius of the circle to be smaller than its steady-state value. The result of this action is evident in the drawing of Fig. 13-1A.

Modified Circuit Using Pentodes

A modification of the circuit of Fig. 13-2 employs pentode tubes instead of triodes. This circuit appears in Fig. 13-3, and, as can be seen, is quite similar to the previous circuit studied. The primary difference between the two circuits is that in the one employing pentode tubes, the signal to be observed is fed into the *screen-grid* circuits rather than into the control-grid circuits. Because of this method of signal introduction, grid capacitors *C2, C3, C4,* and *C5* of Fig. 13-2 are no longer needed, and the center taps of the secondaries of *T1* and *T2* are, therefore, grounded.

You will also notice that a bias battery is no longer being used, the necessary bias being obtained by the voltage drop across cathode resistors *R2* and *R3.* The circuit is balanced in a similar manner to the previous one. That is, with the deflection sensitivities of both sets of deflection plates assumed equal, each amplifier is completely symmetrical with respect to ground, and the components comprising the vertical amplifier duplicate those of the horizontal amplifier. Again, since the deflection sensitivities of the two sets of plates are not quite equal in practice, slight circuit adjustments must be made to obtain equal horizontal and vertical deflection. The same supply voltage is used for the screen and plate circuits, the screens of all tubes being directly connected together, with *R12* serving as a common screen dropping resistor.

By now it should be readily understood that without an input signal applied to the screen grids, a circular trace will be produced upon the screen of the cathode-ray tube. Of course, the resistance *R1* must be equal to the reactance of *C1* at the frequency of operation, to produce two equal 90° out-of-phase signals. The diameter of the trace depends upon both the bias on the tubes and the strength of the time-base signal input.

In order to cause radial deflection of the circular trace, the input signal must modulate the time-base signal and thus produce sum and difference frequencies in the output of both push-pull networks. Again, to achieve this, the input signal must operate over a nonlinear portion of the *plate-current screen-grid voltage* (i_b-e_{sg}) transfer characteristic. In this circuit, it is the *B+* voltage on the screen grid which determines the operating point of the input signal. In order for the signal to operate over the nonlinear portion of

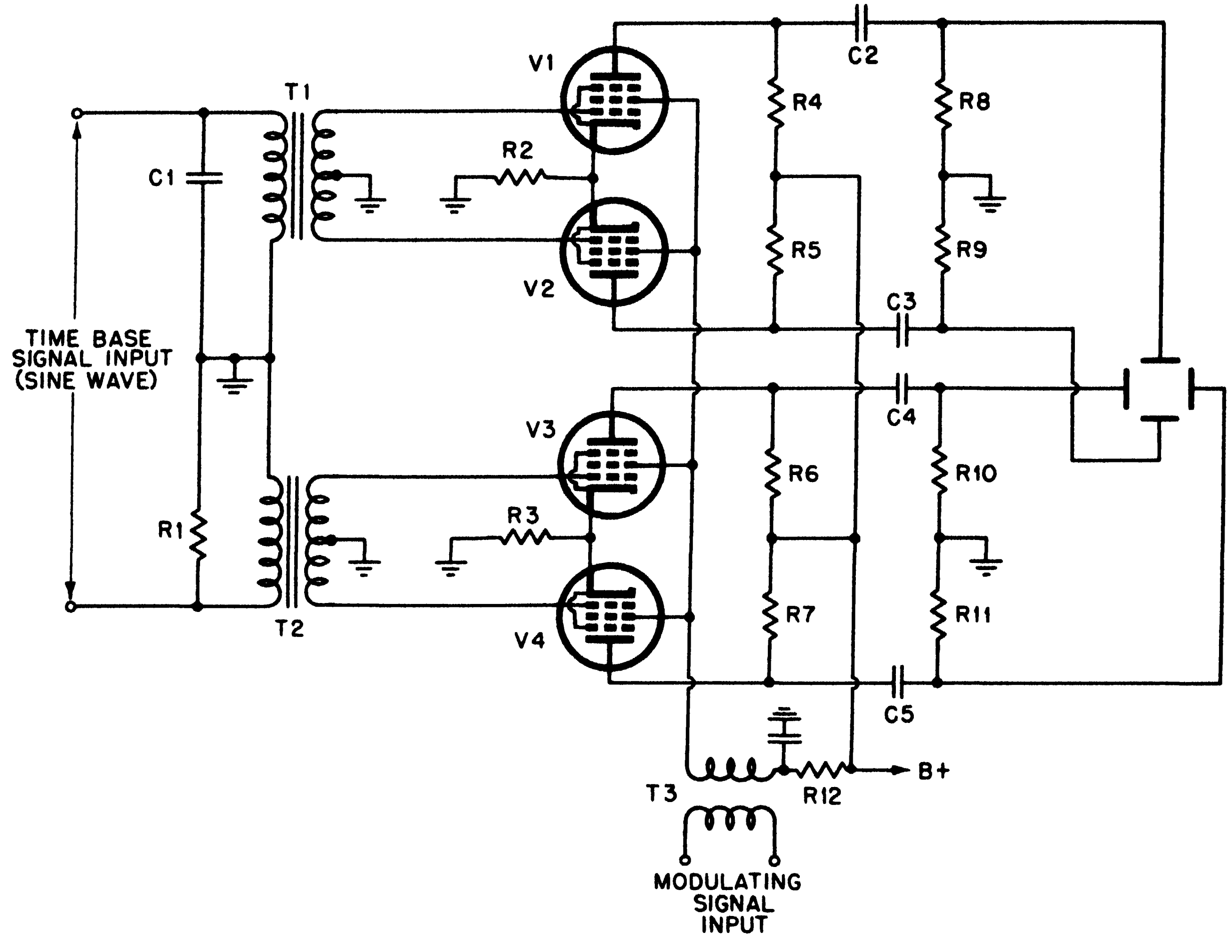

Fig. 13-3.—Modified balanced amplifier circuit using pentodes; the signal to be observed is fed into the screen-grid circuits.

this i_b-e_{sg} characteristic, the screen voltage has to be lower than the plate voltage of the tubes. The exact value of the bias and plate voltage of the tube will determine which curve of a family of i_b-e_{sg} characteristics the modulating signal is to operate over. The *B*+ potential on the screen grids is determined by the value of resistor *R12*. Consequently, we see that the fixed *B*+ on the screen grids in the circuit of Fig. 13-3 plays a similar role to that of the fixed bias in the circuit of Fig. 13-2. Since the value of screen-grid potential controls the flow of plate current, it, therefore, controls the gain of all four amplifiers. In the circuit of Fig. 13-3, the value of the bias is not as critical as in the other circuit.

In the circuit of Fig. 13-3, the modulating signal is seen to be transformer-coupled through *T3* to the screen grids of the tubes. This means that if the modulation signal is sinusoidal, the instantaneous radius of the resulting pattern will decrease when the sine wave goes positive and increase when the sine wave goes negative. What happens is that the modulating signal undergoes a 180° phase reversal due to transformer *T3*. Thus the positive half-cycles of the input signal are negative when they reach the screen grids and, therefore, cause the plate currents of the tubes to decrease because of reduced screen voltages, and the negative half-cycles of modulating signal are positive when they reach the screen grid and thus increase the screen potentials, and hence the plate currents.

If the modulating signal is capacitor-coupled directly to the screen grids, it will not undergo any phase reversal, and the modulation of the trace will then be directly proportional to the amplitude of the modulating signal in the same direction. However, it does not matter whether the positive halves of the sinusoidal modulating signal increase or decrease the plate current, because the reproduced pattern will look the same in either case. This will be true for any modulating signal in which the positive and negative half-cycles are mirror images of each other.

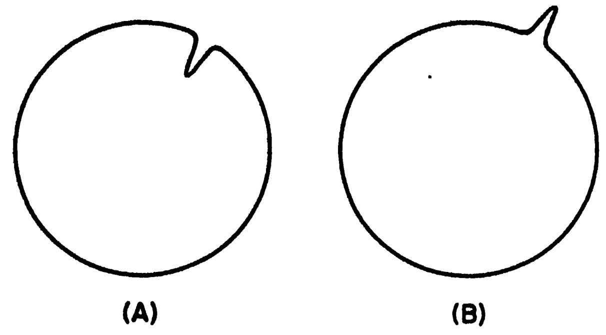

Fig. 13-4.—The effect of feeding a negative pulse into (A) the control grids of the circuit in Fig. 13-2, and (B) the primary of *T3* in Fig. 13-3.

If the modulating signal is a pulse, the situation is different; the polarity of the signal is then important. If a negative pulse is fed directly to the control grids in Fig. 13-2, the plate currents of the tubes will decrease and the circular trace will be deflected *inward*. This is indicated in pattern (A) of Fig. 13-4, where a single pulse is shown deflecting the trace toward the center of the circle. If a negative pulse is applied across the primary of *T3* in Fig. 13-3, then it will appear in the resultant pattern going radially *outward* as indicated in (B) of Fig. 13-4.

When the signal to be observed is fed to the screen grids of the amplifiers, it has to be considerably stronger than when fed to the control grids to produce the same amount of radial deflection. A strong input signal to the screen grids is easily obtained from a stage of preamplification. A simple circuit illustrating this is shown in Fig. 13-5. The modulatng signal input is fed to the grid of the tube via a *C1-R1* coupling arrangement. The bias for the tube is obtained by the voltage drop across the cathode resistor *R2*. Capacitor *C2* serves as a bypass to the signal currents. *R3* is the screen dropping resistor, *C2* is the screen bypass capacitor, and *R4* is the plate load resistor.

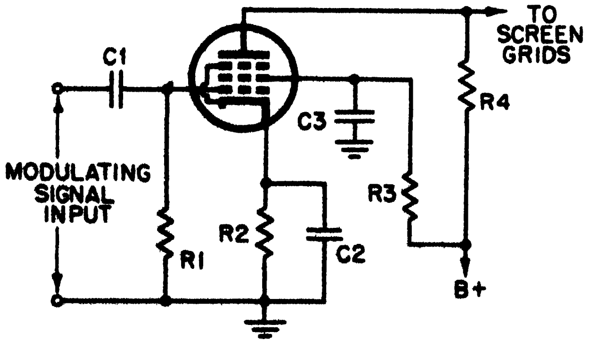

Fig. 13-5.—A preamplifier used to increase the modulating signal strength before it is fed into the screen grids of the modified balanced amplifier circuit of Fig. 13-3.

The screen grids of the circuit of Fig. 13-3 are directly connected to the plate of the amplifier of Fig. 13-5, thus directly coupling the amplified modulating signal. Transformer *T3* and resistor *R12* of Fig. 13-3 are no longer required. Thus, the screen grids of the push-pull amplifiers have the same *B+* voltage as the amplifier of Fig. 13-5. Since there is an inherent 180° phase shift between the control grid and plate circuit of an amplifier, a negative input modulating pulse will appear at the screen grids of the push-pull amplifiers as a positive pulse. In like manner, a positive pulse input will appear as a negative pulse at the screen grids.

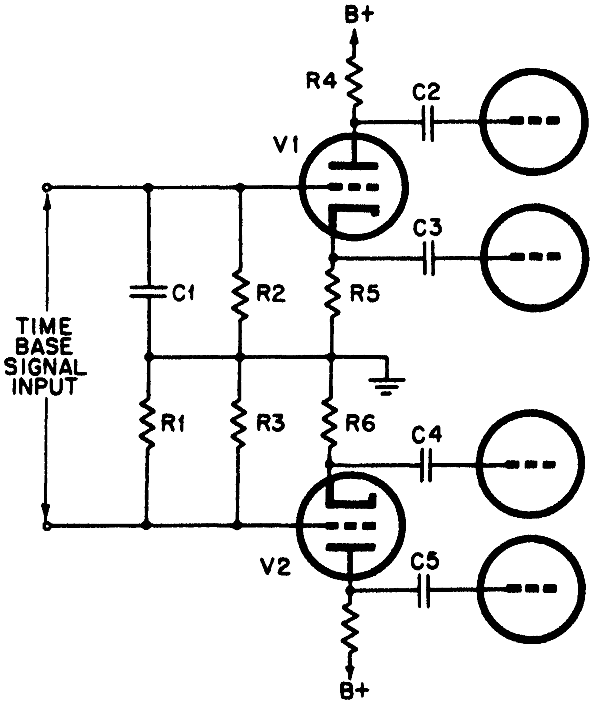

Fig. 13-6.—A simplified phase-inverter circuit for producing a circular sweep.

Instead of using transformers *T1* and *T2* in Fig. 13-3 to impress 180° out-of-phase, but equal amplitude, signals upon each set of push-pull grids, it is possible to employ some form of phase-inverter circuit. A simplified arrangement of such a circuit is shown in Fig. 13-6. Capacitor *C1* and resistor *R1* form the usual phase-splitting circuit with the voltages across each component equal but 90° out of phase. The voltage drop across *C1* is applied to the grid of phase-inversion tube *V1* and the drop across *R1* is applied to the grid of phase-inversion tube *V2*. Since there are no bypass capacitors in the cathode circuits, the signal currents flowing through each tube will cause a voltage drop across both the plate and cathode load resistors. The signal voltage drops across the cathode and plate load resistors of each tube are 180° out of phase with each other, however, due to the normal tube action. By making these resistors of the same value, the signals fed to the grids of each push-pull amplifier system will be opposite in phase, but equal in amplitude. For complete symmetry of the circuit, grid resistors *R2* and *R3* are made equal, as well as load resistors *R4*, *R5*, *R6*, and *R7*. Capacitors *C2*, *C3*, *C4*, and *C5* are also all equal, and serve to couple the signal voltages to the push-pull grids. A phase-inversion system of a different type, however, is used in the circuit of the Du Mont Type 275A Cathode-Ray Polar Co-ordinate Indicator, discussed in Chapter 22.

Radial Deflection by Means of Additional Electrodes

There are two principal methods whereby a circular trace can be radially deflected by use of additional elements inside the cathode-ray tube. Both of these systems have previously been discussed in Chapter 6 with reference to the J scope and the *truncated-cone* scope. The truncated-cone system predates the J scope. In both systems, two sets of deflection plates are used in conjunction with the additional electrodes.

The circular trace is produced by feeding the input signal to any simple 90° phase-splitting network and applying the individual outputs from this network to each set of deflection plates. The signal to be observed which modulates the circular trace is fed to the additional electrodes, whether it be the truncated cones or central terminal on the J scope. What happens is that the modulating signal sets up an additional electrostatic field around these new electrodes, the instantaneous polarity of the signal determining in which direction the beam will be deflected.

Examine Fig. 13-7. Part (A) of this drawing shows a cross section of a J-type cathode-ray tube with the central deflection electrode. With the modulating signal fed between the center electrode and the aquadag coating, it can be readily understood that an electrostatic field will exist between these two elements due to the applied signal. Since the electron beam (which forms the circular trace upon the screen) rotates in the space between the aquadag coating and the central electrode, it will be affected by this new electrostatic field.

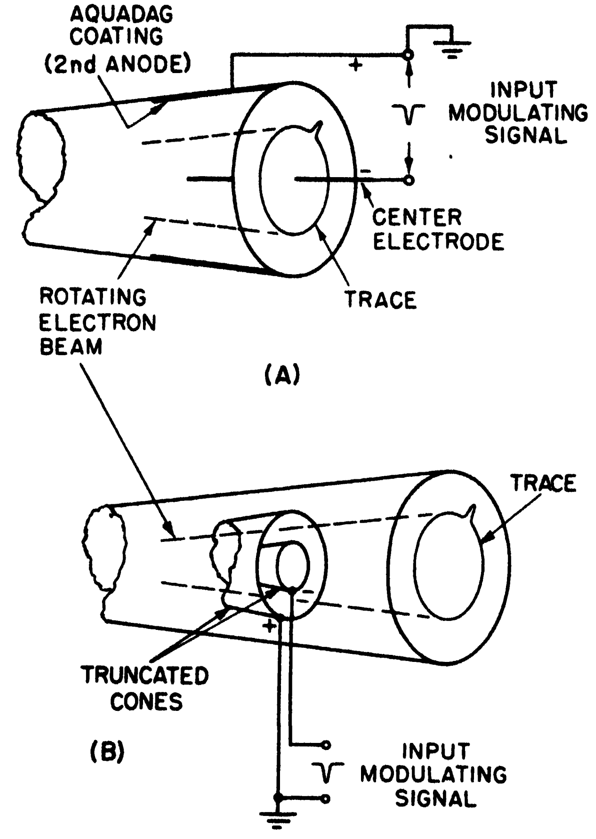

Fig. 13-7.—Radial deflection produced by an additional electrostatic field in the cathode-ray tube using (A) a central electrode and an aquadag coating, and (B) a set of truncated cone deflection plates.

If a negative pulse is fed to these electrodes as indicated in the drawing, the aquadag coating will become positive with respect to the central electrode. Since the beam is in itself a stream of negative electrons, it will be deflected toward the positive electrode, in this case the aquadag coating. This is the same thing as saying that the beam will be deflected away from the negative terminal, which is the central electrode. In other words, a negative input pulse increases the instantaneous radius. This method of operation is usually employed in the radar units where such a tube is used.

In part (B) of Fig. 13-7, we see a cross-sectional view of a typical cathode-ray tube employing truncated cones for radial deflection. One cone is usually connected to the second anode; in the drawing, this electrode is the outside cone which is shown grounded. The electron beam producing the circular trace rotates in a path through the space bounded by these two truncated cones. Thus a signal applied between the two cones will set up an electrostatic field which will act upon the beam. If the input signal is a negative pulse, the inner cone will be negative with respect to the outer one during the pulse period and will repel the beam radially outward. This can be looked upon as either a repelling effect between the negative inner cone and the beam of negative electrons, or an attracting effect between the positive outer cone and the beam of electrons.

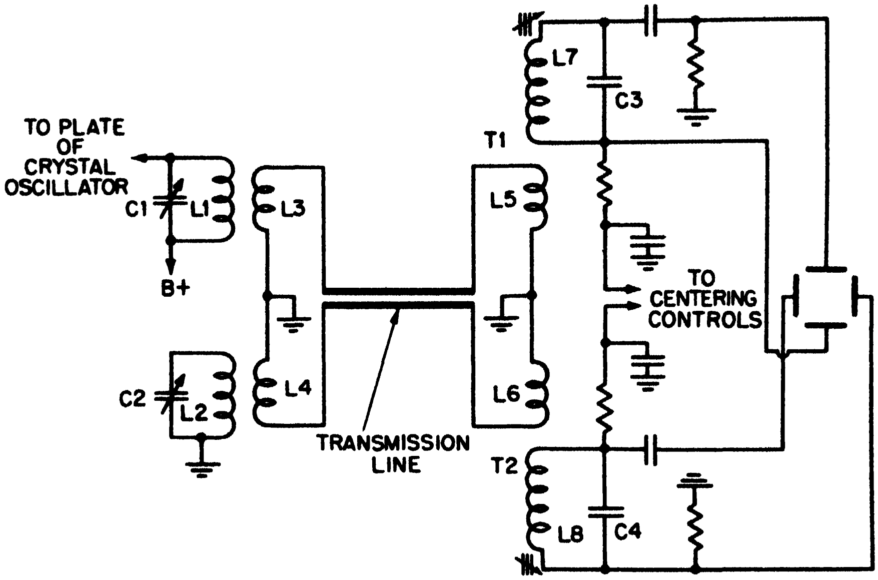

Fig. 13-8.—Circuit used to produce the circular sweep in the J scope of the SCR-584 radar unit of the United States Army.

Phase Splitter Used in the J Scope

In the variety of radar sets in which the J scope has been employed, the two 90° out-of-phase signals necessary to form the basic circular trace were not usually obtained from a simple *R-C* phase-splitting network. The original signal before phase splitting was normally obtained from some crystal oscillator and then transformer-coupled to the phase-shift network. If the unit in which the oscillator was housed was separated from the oscilloscope unit, then some transmission-line coupling arrangement between the two units must be used. When this is the case, we have to consider the problem of impedance matching between the transmission line and the two units.

As an illustrative example, we will examine the basic circuit arrangement used to produce the circular sweep in the J scopes of the SCR-584 radar unit of the U. S. Army. This circuit appears in Fig. 13-8. Inductances *L1* and *L2* are part of a double-tuned transformer, and are loosely coupled to each other. Inductances *L3* and *L4* are two individual coils which are *tightly* coupled to inductances *L1* and *L2,* respectively. Coils *L3* and *L4* are grounded at one end and the other ends are each connected to one lead of a transmission line. For proper operation, the voltages appearing across *L3* and *L4* must be 90° out of phase. Before we discuss the remainder of the circuit, let us see how the 90° out-of-phase voltages are obtained.

The *L1-C1* tuned circuit is connected to the plate of the crystal oscillator. Since the circuit is tuned, it is a pure resistance at its resonant frequency, and hence the current and voltage in this circuit are in phase with each other. This is indicated in the vector diagram of Fig. 13-9, where i_1 is the current vector and e_1 the voltage vector. It is a fact that the voltage induced into a secondary coil is 90° out of phase with the current flowing in the primary. This is true whether or not the coils are tuned. Thus, in the vector diagram, we draw the induced voltage e_2 of the *L2-C2* tuned circuit 90° away from i_1, the primary current vector. Since this latter circuit is also tuned, at resonance its current i_2 is in phase with voltage e_2. From the vector diagram, we see that the voltages in the *L1-C1* and *L2-C2* tuned circuits are 90° out of phase.

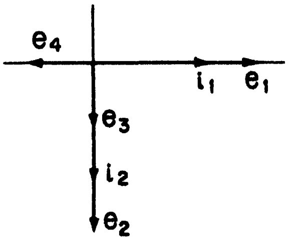

Fig. 13-9.—The phase relations of the voltage and current vectors of *L1, L2, L3,* and *L4* in Fig. 13-8 are indicated here.

Inductance *L3,* being coupled to coil *L1,* will have an induced voltage which is 90° out of phase with the current i_1 in the *L1-C1* network. Likewise, the induced voltage into *L4* from the *L2-C2* tuned circuit will be 90° out of phase with current i_2. This phase relationship is indicated in Fig. 13-9, where vector e_3 is the induced voltage in coil *L3* and vector e_4 is the induced voltage in coil *L4.* From the vector diagram, we find then that the induced voltages e_3 and e_4 which appear across coils *L3* and *L4,* respectively, are 90° out of phase.

Now let us return to the circuit of Fig. 13-8. A transmission line is the connecting medium between the oscillator circuit and the oscilloscope unit, which are both housed on different chassis. The transmission line has a low characteristic impedance; hence, the sweep voltages from the two tuned circuits have to be stepped down for proper impedance matching to the cable. In other words, *L1* and *L3,* and also *L2* and *L4,* are in effect two step-down transformers. The *two* voltages across *L3* and *L4* are coupled by *one* transmission line to coils *L5* and *L6* to couple the signals to the oscilloscope. Note that coils *L5* and *L6* also have one end grounded.

The transfer of these two voltages to coils *L5* and *L6* may not be immediately evident from the circuit of Fig. 13-8. In order to illustrate this voltage transfer, refer to the simplified transmission-line circuit in Fig. 13-10. Since each coil has one side grounded, the cable in conjunction with the ground connections effectively act as two paths of transmission. Thus coil *L3* is in parallel with coil *L5* and coil *L4* is in parallel with coil *L6,* and hence the two 90° out-of-phase voltages also appear across coils *L5* and *L6.*

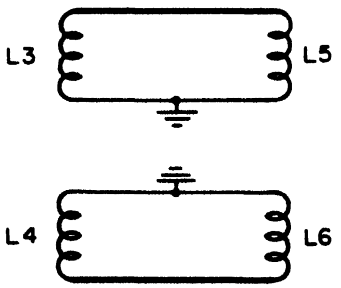

Fig. 13-10.—Simplified transmission-line circuit of Fig. 13-8.

To maintain a continuous impedance match and to supply a high-voltage sweep signal for electrostatic deflection, the voltage output from the transmission line is stepped up. Thus, in Fig. 13-8, *T1* and *T2* are both step-up transformers, with coils *L5* and *L6* having the lower number of turns. The output from these two coils is transformer-coupled to inductances *L7* and *L8,* respectively. The inductances of these latter coils can be varied by iron slug adjustments; *L7* and *L8* form tuned circuits with their respective parallel capacitors, *C3* and *C4.* These tuned circuits are preadjusted to the crystal oscillator frequency.

From our analysis, it is clear that the voltages induced in coils *L7* and *L8* are also 90° out of phase with each other. These voltages are then capacitor-coupled to each set of deflection plates as shown in Fig. 13-8. Consequently, we have the required 90° out-of-phase voltages across the deflection plates.

The degree of coupling between the coils and the symmetry of the circuit are the deciding factors that help make these two voltages in question equal in amplitude so that a *circular* trace, and not an elliptical one, is produced in this radar unit. While we are on the subject of radar, let us see how radial deflection is utilized in many radar systems.

Application of Radial Deflection in Radar Systems

We have mentioned before that the use of a circular trace is one method of enlarging the time baseline. We should remember that the spot travels in the circular path at a *linear rate* of speed. This means that for equal periods of time, the spot traces out the same distance along the circumference for each period. In the majority of radar sets, the 90° phase difference between the two input signals at the deflection plates is such that the spot traces out the circular path in a clockwise direction.

J scopes, when used in radar units, act as *range* scopes. In other words, they are used to determine distances from one point to another by means of radial deflections from the circular path which is calibrated on a linear scale. The exact calibration is dependent upon the rate of spot travel, and hence on the frequency of the sweep signal. Thus, if the frequency of the sweep signal is 10,000 cps, then the spot will trace the circular path 10,000 times in one second. One revolution of spot travel will, therefore, be completed in 1/10,000 of a second, or in 100 microseconds. If the circumference of the circular trace is divided into 20 equal segments, each segment will represent a time lapse of 5 microseconds, and if divided into 40 equal segments, each would represent a time lapse of 2.5 microseconds. With the basic method of calibration known, let us see how it is used to measure distance in radar systems.

A radar signal, which is an electromagnetic wave of very short wavelength, travels at the speed of light, namely at 186,300 miles per second. Thus, if a radar signal travels 18.63 miles, it will have traversed this distance in 18.63/186,300 seconds, or in 100 microseconds. Radar signals have the property of being reflected from metal objects or mountains similar to a sound wave. The reflected signal in either case is commonly known as an *echo.* To determine the distance from some specific target, such as a gun installation or airplane, the directional antenna of the radar set is pointed in the direction of the target, and the system transmits a sharp pulse of very short time duration. (The time duration of the pulse is very much smaller than the time it takes to complete one revolution of the circular trace.) At the instant that this pulse is transmitted, it is also registered on the screen of the tube as a radial deflection along the circular trace at some specific starting point. The design of most radar units is such that the pulse, when it is applied to the central deflection electrode of the J scope, is negative, which assures an outward radial deflection. The pulse will appear similar to that indicated in Fig. 13-11, where it is labeled as the "Main Pulse."

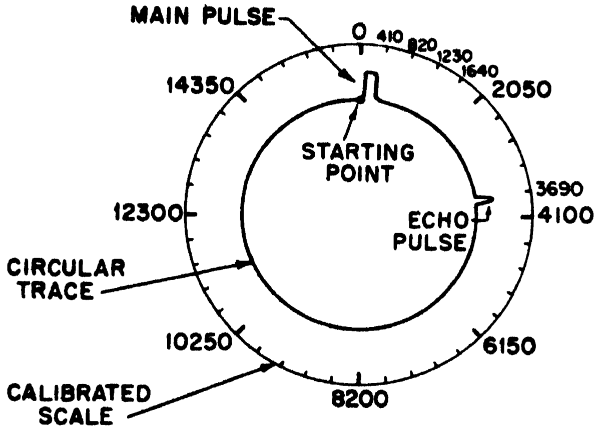

Fig. 13-11.—The distance from the radar transmitter to the reflecting object is obtained directly by reading the clockwise distance from the beginning of the main pulse to the echo pulse on the face of the J scope shown here.

The radar antenna picks up the return echo, which is also in the form of a pulse, and passes it on to the central electrode of the J scope after proper amplification. The echo pulse is fed to this electrode in the same polarity as the main or transmitted pulse so that it will also cause an outward deflection of the circular trace. The time lapse between the transmitted and received, or echo pulse is the time taken by the pulse to travel *to and from* the target.

Let us refer to Fig. 13-11 once again. The time it takes the sweep signal to complete one revolution of the trace is, according to our previous example, 100 microseconds (with a sweep frequency of 10,000 cps). The calibrated scale seen around the outside of the trace is divided into 40 equal divisions so that each represents a time lapse in the spot travel of 2.5 microseconds. For the moment, let us neglect the distance calibration on the outside of this scale. Starting from the main pulse, the echo or received pulse is seen to be picked up nine divisions away from the starting point. This means that 9 × 2.5 or 22.5 microseconds have elapsed between transmission of the main pulse and reception of the echo pulse. Since the distance between pulses is twice the distance to the target, only 11.25 microseconds of time have elapsed in reaching the target.

On the basis of the speed of radar waves, in 11.25 microseconds, a radar wave will travel $186{,}300 \times 11.25 \times 10^{-6}$ miles, or approximately 2.1 miles. Since there are 1,760 yards in a mile, this distance is approximately equal to 3,690 yards. Thus the target is 3,690 yards away from the radar system

Since the scale is supposed to represent distance to the target (i.e., one-half the total distance), it should be calibrated on the basis of 50 microseconds for one full revolution, instead of 100 microseconds. Therefore, instead of using 18.63 miles or 32,800 yards as full-scale calibration of the circular scale, only half, or 16,400 yards, is the actual full-scale calibration employed. Hence, if the received pulse is just about coincident with the transmitted pulse, it took the transmitted pulse approximately 50 microseconds to reach the target 16,400 yards away. The calibrated scale in Fig. 13-11 is, therefore, based on 16,400 yards as the maximum distance for one revolution, each of the 40 equal divisions representing a distance of 410 yards. Therefore, all the radar operator has to do is to read the target distance in yards next to the echo (smaller pulse).

The magnitude of the received or echo pulse indicates whether the antenna is lined up in the direction of the target. Hence, a maximum echo pulse tells us that the antenna is pointed directly at the target.

Two such J scope arrangements can be used together, each with a different sweep frequency and hence with differently calibrated scales. One can be used for coarse distance determination and the other as a vernier scale.

Mechanical Generation of Circular Trace

Another method of forming a circular trace on the cathode-ray tube is to use a special two-phase generator.[2] The output from this generator, if sufficiently large, can be fed directly to the deflection plates of the cathode-ray tube, or it can be fed through an amplifier arrangement similar to those previously discussed. Whatever system is used, the two 90° out-of-phase signals should have the same magnitude in order to produce a circular trace. For radial deflection of the beam, either the J scope, or the truncated cone with the special amplifier arrangements shown in Figs. 13-2, 13-3, or 13-6, may be used.

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[2] The theory of two-phase generators can be found in most good texts on a-c machinery.

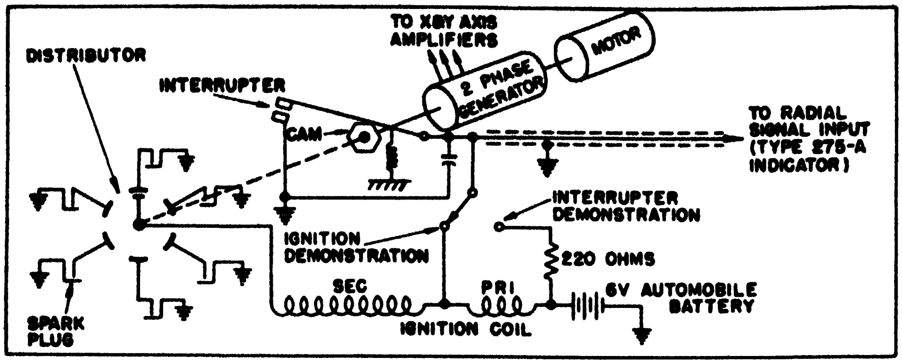

Fig. 13-12.—A two-phase generator produces the circular sweep here used to study the ignition system of a six-cylinder automobile engine.

Courtesy Du Mont Labs.

In the Du Mont type 275A oscilloscope,[3] such a generator is used in conjunction with an amplifier arrangement for radial deflection. Of its many possible uses, the chief application of this unit is the study of all types of rotating machinery. The shaft of the generator is normally coupled to the shaft of the rotating equipment from which the signal to be observed is obtained. The equipment has been used for the testing of automobile and aircraft engines. According to the manufacturer of this unit,[4] timing, chatter, and valve adjustment for each cylinder may be displayed with respect to crankshaft angle. Vibrations may be located, detonation (knock) studied, fuels of different octane ratings distinguished, and carburetor adjustment facilitated.

The crankshaft is coupled to the generator shaft. The generator will thus be driven at the same speed as the engine. The circular trace rotates at the speed of the crankshaft and is, therefore, automatically synchronized. The operation of each cylinder is displayed around the circular trace at approximately equal intervals.

The circuit arrangement of Fig. 13-12 is that employed in the Du Mont laboratory for the analysis of the ignition system of a six-cylinder automobile engine. This laboratory setup duplicates the ignition system and permits a study of interrupter adjustment and operation and also a check on spark-plug firing and electrode spacing. An electric motor was used to drive the interrupter, distributor, and the two-phase time-base generator. For operation in an automobile, this generator may be coupled to the accessory shaft, either directly or through an auxiliary tachometer shaft.

The performance of the interrupter may be determined from a study of Fig. 13-13. These oscillograms show a pulse of voltage generated each time that the contacts were opened by the cam. In general, the oscillogram is not a simple square-pulse pattern because of the inductance of the coil and the capacitance across the interrupter. The correct setting of the contacts can be accurately determined because the effect of each adjustment can be observed on the cathode-ray tube. Part (A) of Fig. 13-13 shows the appearance of the oscilloscope pattern when the distributor contacts are correctly adjusted. When they are slightly out of adjustment, the pattern will appear as shown in part (B). In this pattern, a slight contact bounce is evident from the series of uneven oscillations in each trace. A greater misadjustment will produce a more pronounced bouncing effect, as shown in pattern (C).

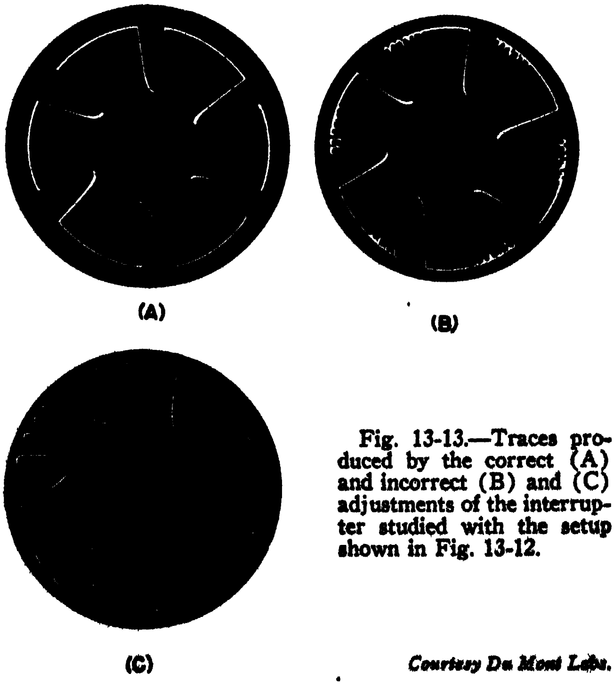
(A) (B) (C)

Fig. 13-13.—Traces produced by the correct (A) and incorrect (B) and (C) adjustments of the interrupter studied with the setup shown in Fig. 13-12.

Courtesy Du Mont Labs.

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[3] The complete schematic and circuit description of this unit is included in Chapter 22.

[4] Marion, M., and Kline, M. B., "The Du Mont type 275-A cathode-ray polar-coordinate indicator," *The Oscillographer,* vol. 9, no. 5, September and October, 1947, copyright Allen B. Du Mont Laboratories, Inc.; this article gives a detailed description of the unit and its applications.

The performance of the ignition system may be observed if the "Ignition Demonstration" switch is at the position shown in Fig. 13-12; in this setting, the "Radial Signal Input" terminals are connected across the primary winding of the ignition coil.

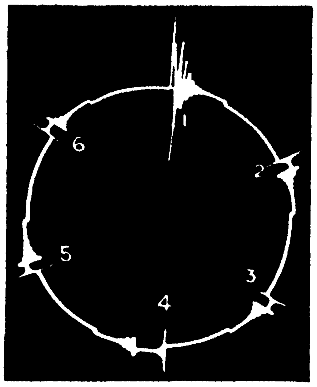

Fig. 13-14. — Oscillogram containing the traces of each of the six spark plugs in a six-cylinder engine.

Courtesy Du Mont Labs.

Fig. 13-14 is a composite oscillogram containing six separate traces, one for each spark plug in the ignition system of a six-cylinder automobile engine. Each trace is an oscillogram of the shock oscillation voltage across the primary winding of the ignition coil during the interval that the high secondary voltage is applied to a particular spark plug. From these traces, it is possible to determine whether a plug is firing, the length of the firing period, etc., and consequently, the electrode spacing and condition of the corresponding plug.

The trace for the first plug in Fig. 13-14 represents a plug that fails to fire because the electrodes are spaced too far apart to permit breakdown. Inspection of the oscillograms for plugs *2, 3, 5,* and *6* reveals that the amplitudes of the traces suddenly diminish, that is, they are quickly damped out. The exponential decrease in voltage is due to the loading of the coil and is evidence that these plugs are firing normally.

Because the spark plug and the gap in the distributor are in series, the applied voltage is across the distributor when a plug is shorted. A shorted plug may be identified by an oscillogram such as that shown for plug *4* in Fig. 13-14; the indicated breakdown period is longer than that for a plug which is firing normally. This occurs because a voltage below the minimum required for the maintenance of the spark between the two gaps in series is sufficient to sustain a spark across the distributor alone. Thus defective spark plugs may be quickly located using this technique, avoiding the usual cut-and-try methods. Similarly, this method may be applied to the timing and adjustment of any type of rotating machine.

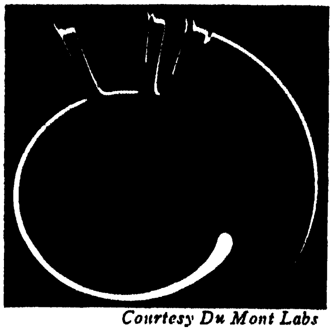

Fig. 13-15.—Trace of relay performance obtained with an oscilloscope using a two-phase generator driven by a synchronous motor.

Courtesy Du Mont Labs

Relay performance can also be checked by means of this method. The photo appearing in Fig. 13-15 shows the bounce of the moving arm of a relay directly after activation. The small semicircle at the left of the figure represents the time before activation. Radial deflections at the top of the figure indicate bouncing at the time of activation. The larger semicircle at right represents the rest period of the activated relay arm after bouncing had subsided. This oscillogram was obtained by driving the two-phase generator with a synchronous motor, and applying a d-c potential through the relay contacts to the radial input of the oscilloscope.

Spiral Time Bases

Sometimes the linear, and even the circular, time base may not be long enough to record certain data. Also, due to the special nature of the phenomena observed, the circular time base may not produce the required accuracy of measurement. The *spiral time bases* fill the need for longer time bases and are convenient for use in radar and industrial applications.

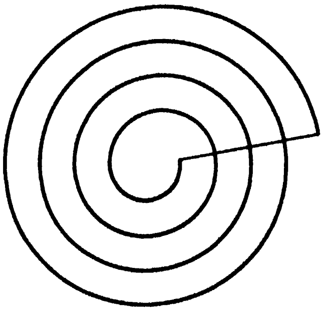

Fig. 13-16. — An ideal spiral time base showing four revolutions.

Spiral time bases are evolved from circular time bases. Fig. 13-16 illustrates an ideal spiral time base which has undergone four revolutions or turns. The length of such time bases can be varied easily by simply changing the number of turns of the spiral. The beam spot rotates at a constant rate of speed so that equal divisions along the spiral represent the equal increments of time. Let us see how such time bases are formed.

As there are a number of different methods of producing a circular time base, there are also various ways of producing a spiral time base. For the moment, assume that a circular sweep signal has been produced in the conventional manner and the trace is displayed upon the screen. The beam spot continuously traces out the circular path as long as the sweep signal is applied. The diameter of the circular trace depends upon a number of factors, such as the amplitude of the applied sweep signal and the gain of the amplifiers to which the sweep signal is fed. If the amplitude of the sine wave causing the circular sweep signal or gain of the amplifiers is varied in a *linear* manner, then the spot, instead of tracing out a single circle, will trace out a spiral, provided the circular speed of the beam is considerably greater than the linear speed. The linear variation of the amplitude, and hence the diameter of the circular time base is easily obtained by use of a linear time-base generator or a sawtooth potentiometer. The generator can be any one of the number of forms (see Chapter 9) that will produce a sawtooth voltage output.

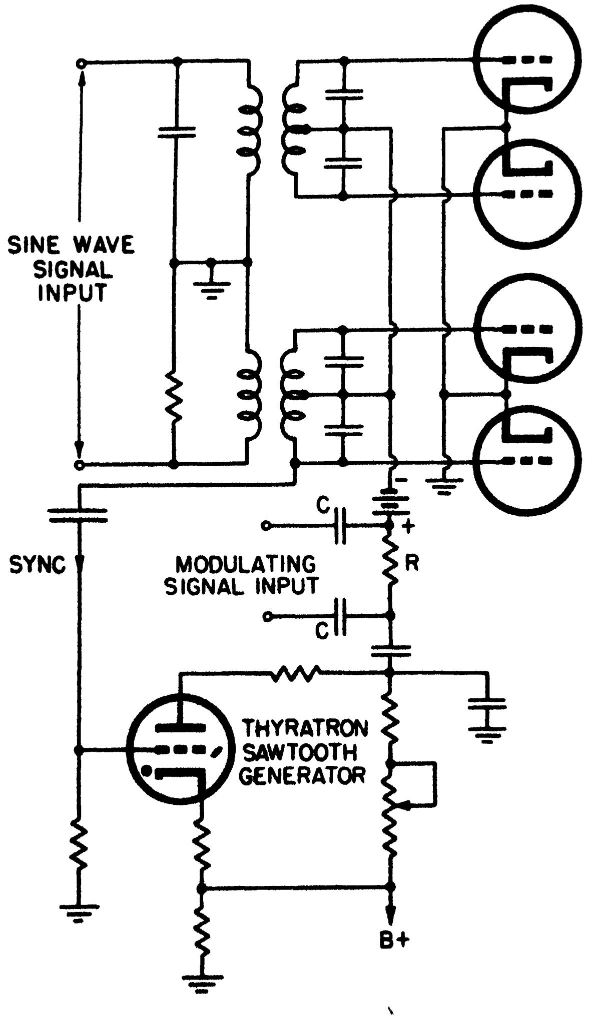

Fig. 13-17.—Simplified circuit for producing a radial deflection for a spiral sweep.

The input circuit of Fig. 13-2 is repeated in Fig. 13-17, but in the latter circuit, a thyratron sawtooth generator is used to vary the gain of the two sets of push-pull amplifiers by changing the bias on the grids. Note that the sawtooth signal is fed to the grid in series with the battery voltage and the modulating signal input. Part of the sine-wave signal input is fed to the grid of the thyratron to insure synchronization of the sawtooth with the sine-wave time-base signal, so that a stationary pattern appears upon the screen.

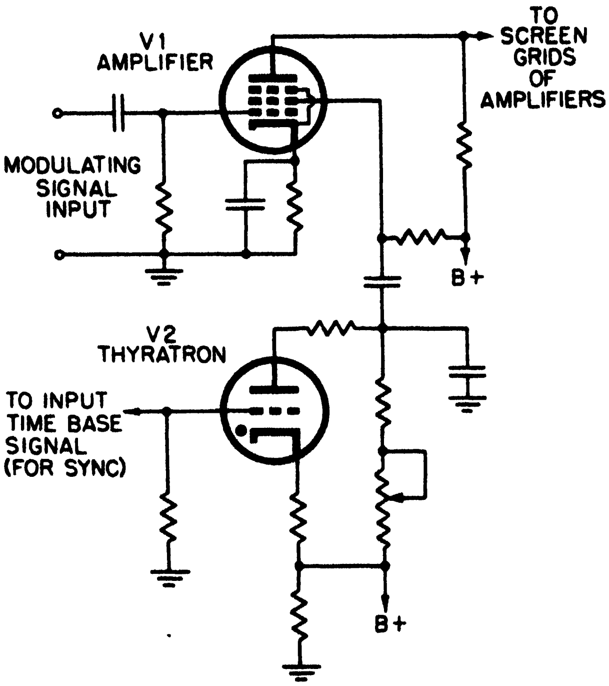

Fig. 13-18.—Circuit for obtaining a spiral sweep using pentode amplifiers.

The gain of the amplifiers can also be varied linearly by applying a sawtooth signal to the screen grids of the push-pull amplifiers. The circuit arrangement of Fig. 13-18 illustrates this. Tube *V1* is the same amplifier as that in Fig. 13-5, which is used to amplify the modulating signal before it is applied to the screen grids of the push-pull amplifiers. Tube *V2* is a thyratron sawtooth generator which is used to vary the gain of *V1* and hence the gain of the push-pull amplifiers. The gain of *V1* is varied by applying the sawtooth signal to its screen grid.

Let us examine the circuits shown in Figs. 13-17 and 13-18, and see how the spiral time base is formed. For the moment, let us assume that the modulating signal is not yet applied, and that the frequency of the applied time-base sine wave is four times that of the sawtooth signal. This means that four turns of a spiral trace will appear on the screen of the tube. Refer to Fig. 13-19. Parts (A) and (B) represent the sine-wave and sawtooth signal, respectively, while (C) shows the shape of the resultant signal. Two such signals, 90° out of phase, appear across each set of deflection plates. The dashed lines of part (C) indicates a linear increase in the sine wave. The four resulting spiral turns are shown in part (D). Since the amplitude of the sawtooth *increases* linearly with time (hence decreasing the bias on the time-base generator and increasing its gain), the spiral time base will open outward from the center of the spiral. The radius of the first spiral turn depends upon the minimum gain of the amplifiers. If the frequency of the sine wave is not an integral multiple of the sawtooth, the spiral will not remain stationary, but will revolve.

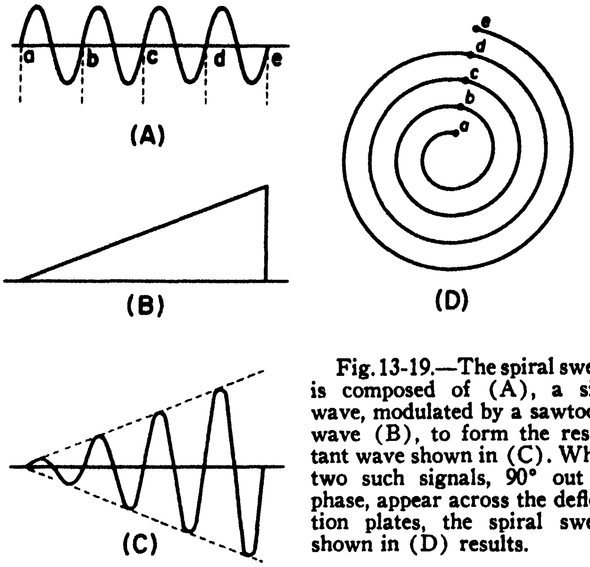

Fig. 13-19.—The spiral sweep is composed of (A), a sine wave, modulated by a sawtooth wave (B), to form the resultant wave shown in (C). When two such signals, 90° out of phase, appear across the deflection plates, the spiral sweep shown in (D) results.

The rotation of the spot in tracing out the spiral time base depends upon the relative angle of lead or lag between the two 90° out-of-phase sine waves at the deflection plates. In the illustration of Fig. 13-19D, the spot is shown as traveling counterclockwise. During the first cycle of the sine wave of Fig. 13-19, time *a* to *b*, the amplitude of the sawtooth increases, and the first turn of the spiral is traced out from point *a* to *b*. During the next cycle of the sine wave, the spiral is traced out from point *b* to *c*, during the third cycle from *c* to *d*, and for the fourth cycle from *d* to *e*. Because the sawtooth wave is linear, the radial separations between spiral turns are all equal. Although not shown in Fig. 13-19D, if not blanked out, a retrace would appear from point *e* to *a*, provided it occupies negligible time in comparison with the complete cycle of the sawtooth wave. Such an ideal retrace was indicated in Fig. 13-16.

When a modulating signal is applied to the circuit shown either in Fig. 13-17 or in Fig. 13-18, the spiral trace will be radially deflected. The position of the radial deflection along the spiral is dependent upon the time elapsed between the sweep initiation and the application of the modulating signal. For the circuits under discussion, the modulation voltage applied to the deflection plates will be increased along with the gain of the amplifiers. This means that the amplitude of a pulse will be progressively increased along each turn of the (outgoing) spiral.

This is best indicated in Fig. 13-20. Pulses *a, b,* and *c* represent a periodic signal which has the same repetition frequency as the input sine wave. This means that one pulse will appear along each turn of the spiral along the same radial line. The actual amplitude of the input pulse is constant, but, as is seen from Fig. 13-20, pulses *a, b,* and *c* on the oscilloscope have different amplitudes. These differences in amplitudes are in proportion to the gain of the amplifiers at that instant. If point O represents the center of the spiral, then the radius of the first spiral turn is indicated by dimension OA, the radius of the second spiral turn by OB, and that of the third by OC. The proportional relationship between the radii and the pulse amplitudes is expressed as follows

$$\frac{OA}{\text{height of pulse } a}=\frac{OB}{\text{height of pulse } b}=\frac{OC}{\text{height of pulse } c}.$$

This is evident from the drawing.

The discussion thus far was limited to the case where a sawtooth signal varies the gain of the amplifiers, thus producing the spiral time base. If the amplitude of the input sine wave is varied linearly, *before* it is phase-split, then it is possible to use a J scope to produce the

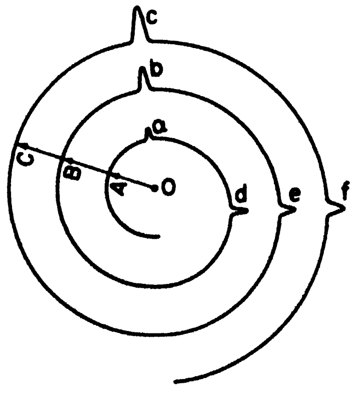

Fig. 13-20. — Radial deflection of a spiral trace by a periodic pulse signal whose frequency is equal to that of the input sine wave. Pulses *a*, *b*, and *c* derive from a circuit in which the gain of the amplifiers is varied linearly; *d*, *e*, and *f* from a circuit in which the sine-wave input is varied before amplification.

spiral trace, in conjunction with the amplifier system. Whatever system is used, if the amplitude of the sine-wave time-base signal is varied linearly, *before* amplification, then the modulating signal will cause *the same amount of deflection*, regardless on which spiral turn it appears. Thus a periodic pulse with the same repetition rate as the frequency of the input sine wave will produce equal-amplitude pulse deflections along each spiral turn. This is indicated by pulses *d, e,* and *f,* in Fig. 13-20.

Spiral Time Base Produced by Damped Oscillations

In the spiral time bases dicussed thus far, the spiral opened outwardly because of the increasing amplitude of the sawtooth signal. If the applied sawtooth signal *decreased* linearly, then the resultant signal applied to the deflection plates would be a damped sine wave. This is shown in Fig. 13-21, where part (A) represents four cycles of the sine wave, part (B) the sawtooth signal decreasing in amplitude, and part (C) the resultant damped sine wave. With the shape of the sine wave as shown in (C), the spot in tracing out the spiral time base will rotate *inward,* as indicated by the labeling *a* to *e* in part (D) of Fig. 13-21.

From this brief analysis, we surmise that a damped *oscillatory* sine wave can be used to produce a spiral time base. Since the amplitude of damped oscillations decreases in exponential fashion, the spacing between turns of the resulting spiral will not be equal, but will decrease in exponential fashion. A spiral of this type is known as an *exponential* or *logarithmic spiral.* In Fig. 13-22A appears a six-cycle exponentially damped sine wave. Part (B) shows the resulting exponential spiral, which is seen to have six turns. The letters from *a* to *g* on the spiral correspond with those on the damped sine wave.

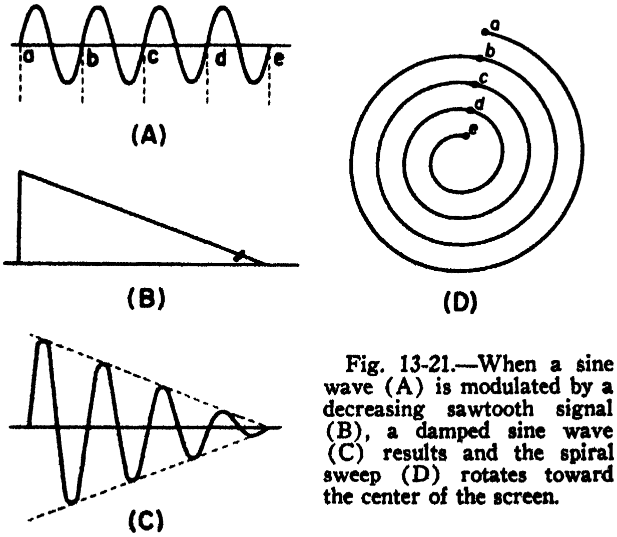

Fig. 13-21.—When a sine wave (A) is modulated by a decreasing sawtooth signal (B), a damped sine wave (C) results and the spiral sweep (D) rotates toward the center of the screen.

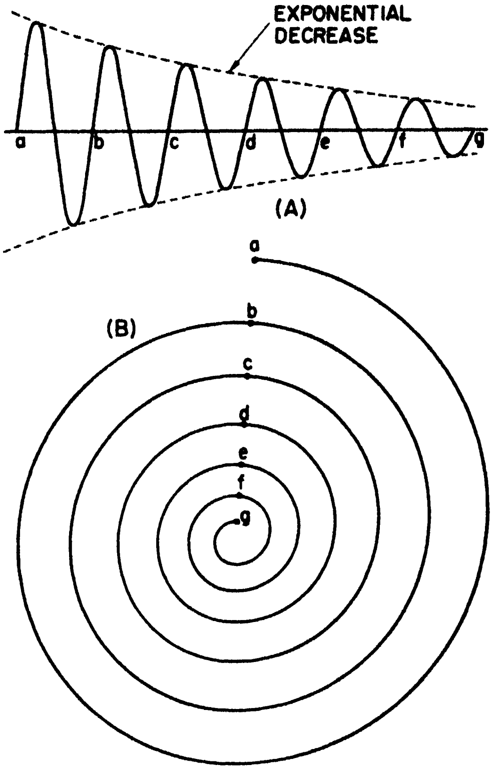

Fig. 13-22.—The exponential spiral shown in (B) results from an exponentially damped sine wave (A). The points *a* to *g* on the spiral correspond to points *a* to *g* on the sine wave.

Such a damped signal can be produced by using an exponential generator (such as a nonlinear sawtooth generator) in place of the linear generator of Figs. 13-17 and 13-18. Another method of producing such a damped signal is to shock-excite an *R-L-C* oscillator circuit and use this oscillatory signal as the input signal to the phase-splitting network of the deflection system. As far as we know, such tests were first conducted in England by Rawcliffe.[5]

The simplified circuit of this arrangement appears in Fig. 13-23. Components *R* and *C* are the phase-splitting network and *R1-C1-L1* is the oscillatory circuit that is in parallel with the phase splitter. A linear time-base generator, such as a thyratron or multivibrator, is used to excite the circuit into oscillations. Resistance *R1* determines the degree of damping of the circuit. Since the oscillations are initially caused by excitation from the linear generator, the spiral and linear base will be in synchronism at all times. The frequency of the oscillatory circuit should be several times that of the linear generator in order for a number of

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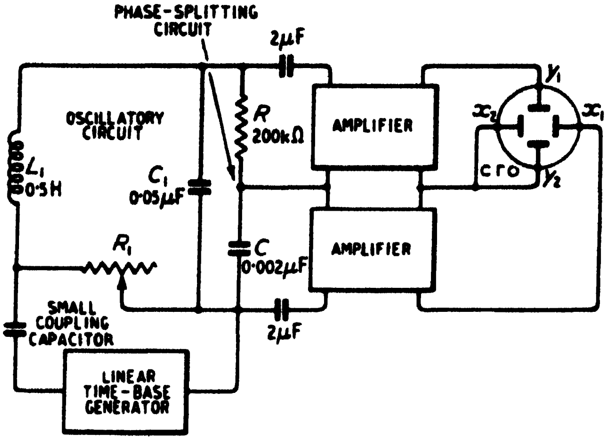
[5] Rawcliffe, G. H., "Shock-impulsed spiral time base," *Wireless Engineer,* vol. 26, p. 242, July, 1949.



Courtesy Wireless Engineer

Fig. 13-23.—A simplified version of a circuit for producing an exponential sweep.

spiral turns to appear. The interesting thing about such an arrangement is that the spiral will remain stationary no matter what the ratio of these two frequencies. However, the shape of the spiral changes with variation in ratio. If the frequency of the linear generator is increased, the center turns of the spiral appear to unwrap themselves with fewer spiral turns resulting.

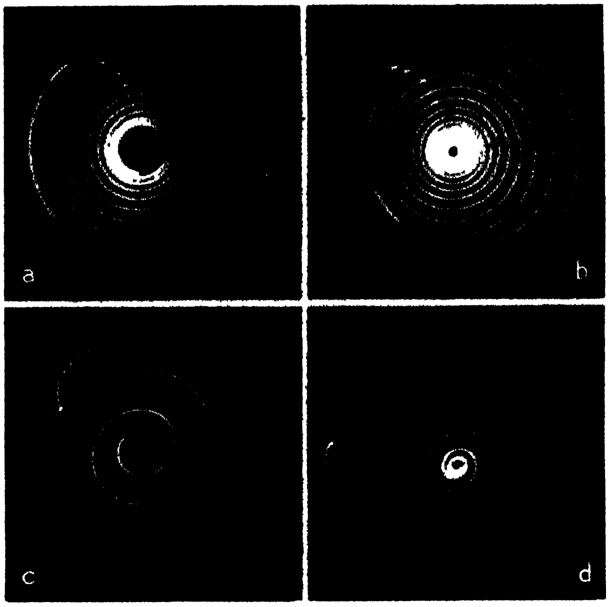
In order that the phase-splitting *R-C* circuit of Fig. 13-23 shall have a negligible effect on the oscillatory circuit, it is necessary to make the impedance of the *R-C* circuit several times that of the capacitive reactance in the oscillatory circuit. The circuit values given are merely examples; a wide variety of values could be chosen.

In Fig. 13-24 appear four oscillograms of spiral time bases taken by Rawcliffe and obtained from the circuit of Fig. 13-23 for different Q values of the oscillatory circuit and various frequencies of the linear generator. The measured circuit Q*'s* and the calculated oscillatory frequencies (designated f) for each pattern were as follows: For pattern (a), $Q = 17.1$ and $f = 2{,}280$ cps; for pattern (b), $Q = 16$ and $f = 1{,}950$ cps; for pattern (c), $Q = 13.4$ and $f = 1{,}040$ cps; and for pattern (d), $Q = 8$ and $f = 780$ cps. The smaller the Q of the circuit, the greater the degree of damping, and hence the greater the difference in radial separation between each successive pair of spiral turns. Thus spirals (a) and (b) of Fig. 13-24, which have higher Q*'s* than the spirals shown in (c) and (d), are more evenly spaced than the latter two. In fact, the Q of the circuit can be determined by certain measurements of the spiral. According to Rawcliffe

$$Q = \frac{\text{length of spiral between two radii}}{2 \times \text{difference between these radii}}.$$

Consequently, this method can be used as a means of determining the Q of certain oscillator circuits.[6]

It is difficult to detect the number of turns in spiral (b) of Fig. 13-24; for spirals (a), (c), and (d), the number of turns are approximately 9½, 3, and 5, respectively, taking into account the time of the retrace. From the number of spiral turns and the known frequency of the oscillatory circuit, we can calculate the approximate frequency of the linear generator used for excitation. For spiral (a), $f = 2{,}280$ cps and the number of turns is approximately 9½, so that the frequency of the linear generator is 2,280/9.5 or 240 cps. In like manner, for spiral (c), the oscillatory frequency is 1,040 cps and the number of turns about 3, so that the linear frequency is 1,040/3 or about 347 cps, and for spiral (d), the linear frequency is 780/5 or 156 cps.



Courtesy Wireless Engineer

Fig. 13-24.—Four spiral sweeps obtained by using the circuit in Fig. 13-23 with various Q values for the oscillatory circuit, and with various linear generator frequencies.

The discussion of nonlinear sweeps thus far was based on cathode-ray tubes of the electrostatic type. By proper design of certain circuit constants, the time bases studied in this chapter can also be formed with cathode-ray tubes employing magnetic deflection.

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[6] Since it is difficult to determine the length of the spiral along the curve, an auxiliary method was devised by the authors for determining the Q from the oscillogram. Using this method, $Q = \dfrac{1.37n}{\log r_2/r_1}$, where r_1 is the radius to some turn of the spiral, r_2 is the radius to another turn of the spiral, further away from the center and *measured along the same radius vector as* r_1 (i.e., $r_2 > r_1$), and n is the number of turns completed in going along the spiral from r_1 to r_2. Hence, if the radii of two *adjacent* spiral turns are measured, n equals 1.

From the standpoint of simplicity of design and ease of operation, however, the electrostatic type of deflection is usually more desirable for producing these nonlinear time bases.

Radial Time Base

If the frequency of the sawtooth generator in the spiral time-base circuits of Figs. 13-17 and 13-18 is made appreciably greater than the frequency of the input sine wave, it is possible to obtain a *radial time base,* rather than a spiral one. If the gain of the amplifiers in question is varied from zero to a maximum by this linear sawtooth wave, then the resulting pattern (when the sawtooth frequency is greater) will appear as the spokes of a wheel similar to the ideal radial time-base pattern shown in Fig. 13-25. If the frequency of the linear generator is not much greater than that of the sine wave, the spokes of the wheel will have some curvature. Let us analyze how a radial time base is formed.

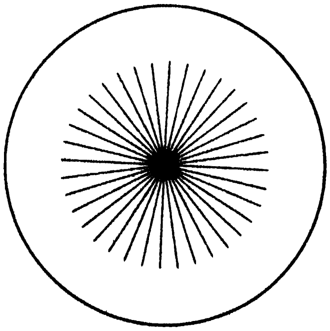


Fig. 13-25.—Radial time base obtained by modulating a sine-wave signal during amplification with a sawtooth-wave signal whose frequency is appreciably greater than that of the sine wave.

Assume that the sawtooth wave changes the gain of the amplifiers from zero to a maximum, that its frequency is 10,000 cps and is 20 times that of the sine-wave input. Considering the sine-wave input alone, a circle will appear upon the screen, as seen in Fig. 13-26A, and since the frequency of the sine wave is 500 cps, one circular revolution of the spot will be completed in 2,000 microseconds. If only the linear sawtooth wave is considered, a *single* cycle of this signal would appear on the screen of the tube, because the same sawtooth signal is applied to both sets of deflection plates. One cycle of the sawtooth wave will be completed in 100 microseconds or in 1/20 of the time to complete one revolution of the circular sweep. If the retrace of the sawtooth is blanked out, only a single diagonal line would appear upon the screen. If the circuits are so adjusted that the start of the sawtooth wave is from the center of the screen, the pattern would appear as shown in Fig. 13-26B.

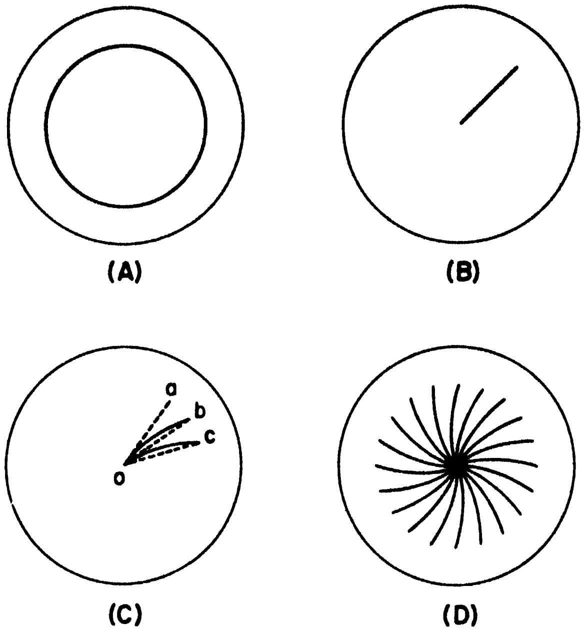


Fig. 13-26.—The formation of a radial sweep. The circular trace due to the sine wave acting alone (A), when combined with the straight-line path of the sawtooth acting alone (B), results in the slightly curving trace shown partially in (C) and completely in (D).

With the basic understanding of how each applied signal acts alone, let us now consider the effect of combining both signals. The spot, by just the action of the sawtooth signal, would normally travel radially from the center of the screen in a straight-line path as indicated by the dashed line *oa* in Fig. 13-26C. However, as the spot starts its travel from point *o,* a circular motion is also imparted to the beam and the spot travels in a parabolic trajectory, ending up at point *b.* The time that it takes the spot to travel the solid line from *o* to *b* is 100 microseconds, in which time it will also travel 1/20 of the circumference of the circle due to the circular sweep; hence the arc subtended by *ab* is 1/20 of the circumference of the circle. As the spot reaches point *b,* the very rapid retrace time of the sawtooth causes the beam to jump back to the center of the screen (point *o*) in a practically straight-line path, as indicated by a second curved radial line *oc* shown in Fig. 13-26C. With the retrace of the sawtooth blanked out, only the curved portions will be visible. After 2,000 microseconds have passed, the sawtooth signal will have undergone 20 cyclic variations, while the sine wave will have passed through only one cycle, and hence 20 individual traces will appear upon the screen [part (D) of Fig. 13-26].

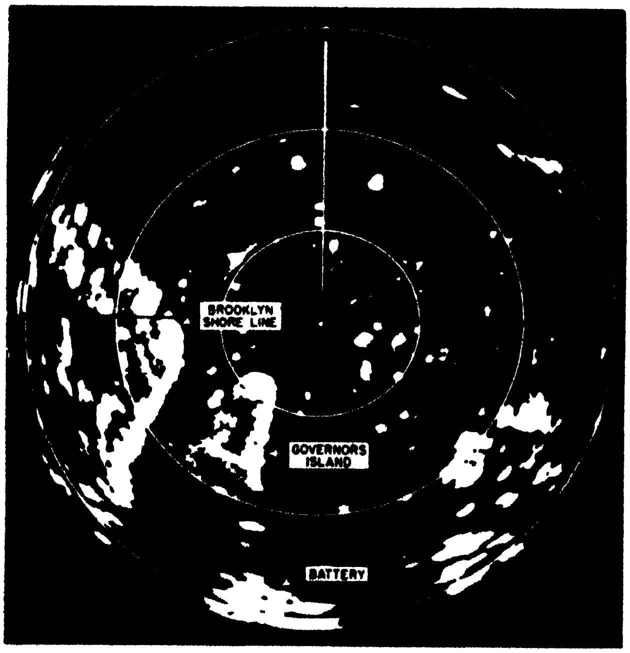
In order to make the traces much more linear, the frequency of the sawtooth, as mentioned previously, has to be much greater than that of the sine wave. The greater the frequency ratio, the better will be the linearity and the greater the number of spokes. The increased frequency of the sawtooth, or decreased frequency of the sine wave, means that, during each spoke trace, the amount of time the sine wave is active is much less than before, thereby reducing the curvature of the trace. If the sawtooth frequency is several hundred times greater than that of the sine wave, the spokes would become effectively linear and appear to merge with one another.

The signal that is to be examined on a radial time base is usually applied to the *control grid* of the cathode-ray tube to intensity-modulate the trace. Thus, if the signal to be observed is strong enough, it will cause certain parts of the trace to be much brighter than others. If a pulse is applied to the grid, it will appear as a bright spot somewhere along the trace. The traces of such patterns can easily be used to measure distance in a manner similar to that described for the J scope. Thus, if the sawtooth sweep has a frequency of 10,000 cps and a single trace or spoke is one inch long as seen on the cathode-ray tube, it will be completed in 100 microseconds, and hence each 0.1 inch would represent a time lapse of 10 microseconds. Thus, if a transmitted pulse is recorded at the start of the single trace and if a return echo from a target shows a bright spot 0.5 inch away from the center of the trace, 50 microseconds have elapsed. The distance to the target is, therefore, reached in half this time or in 25 microseconds, which means the target distance is approximately 8,200 yards away, basing the speed of the pulse travel at 186,300 miles per second.

In many radar applications, radial time bases are used, but they are usually produced in a manner different from that just discussed. In one method, mentioned in Chapter 6, the radial time base is produced by using electromagnetic deflection and rotating the deflection yoke around the neck of the tube at a constant rate of speed. This is one of the many types of PPI (Plan Position Indication) systems used during the war. The voltages applied to the two sets of deflection coils are such as to produce sawtooth current waves flowing through the coils, and hence cause a linear deflection of the beam. The d-c component of the sawtooth wave is adjusted so that the beam describes a path starting out from the center of the screen. If the retrace of the sawtooth is blanked out and if the yoke remains stationary, a straight line similar to Fig. 13-26B will appear upon the screen. When the yoke assembly is rotated, the spot describes the same path as in parts (C) and (D) of Fig. 13-26. If the frequency of the applied voltage to the deflection coils, and hence the sawtooth current, is much greater than the rotation of the deflection yoke, the radial time-base deflections will be almost linear and will fill out the entire screen of the cathode-ray tube.

When used in radar applications; a rotating antenna is mechanically synchronized with the yoke (usually by a servomechanism arrangement), so that the position of the yoke in its circular path will directly indicate the azimuth direction of the antenna. Such a system will display two types of information upon the screen, namely target distance (range) and target direction (azimuth).

In radar practice, the voltage to the deflection yoke is applied only during the transmission of a pulse. The bias arrangement on the grid of the cathode-ray tube is such that during the transmission of pulses the grid is cut off, so that no trace appears on the screen. When a transmitted pulse hits the target, the return echo is picked up by the antenna, amplified, and then applied to the grid of the cathode-ray tube as a positive pulse. This amplified echo pulse is strong enough to drive the grid above cutoff, thereby causing a spot to appear upon the screen. This kind of system will present a map of the area (white on a darker background), scanned by the antenna. A typical PPI pattern of New York harbor at a range of 1½ miles is shown in Fig. 13-27. With suitable calibration of the cathode-ray-tube face, the direction of range of any part of the map becomes known.



Courtesy Radiomarine Corp.

Fig. 13-27.—A typical PPI pattern of New York harbor at a range of 1½ miles.

After the pulse is transmitted, the antenna will rotate a certain distance before the return echo is recorded upon the screen. Due to this fact, it may be thought that the recorded echo will not represent the true direction of the target. This is theoretically correct, but if we realize that the rotation of the antenna is very slow compared to speed of the pulse, it becomes evident that the amount of antenna rotation during the time of pulse transmission and return is negligible. For example, if the target is about three miles away, it will take a pulse approximately 32 microseconds to travel to and from the target (at 186,300 miles per second). If the antenna rotates at the speed of five revolutions per minute, or 30° per second, in 32 microseconds the antenna would have rotated only $30° \times 32 \times 10^{-6} = 0.00096°$, or approximately 0.001°. It is, therefore, apparent that the amount of antenna rotation during the time between a transmitted and received pulse is negligible.

There are other methods besides the rotating deflection yoke whereby such a PPI scan may be produced. One common method employs a fixed-yoke system. Linear sawtooth currents of the same frequency are made to flow through each set of deflection coils so that two mutually perpendicular forces act upon the beam, causing it to take the vector position of these two forces or radially outward. In order to produce the rotation of the radius, the amplitudes of the deflection sawtooth currents are varied sinusoidally at the same rate, while simultaneously maintaining a 90° phase difference between them. The frequency of sinusoidal variation is made much smaller than the frequency of the sawtooth currents to insure linearity.

In radar applications, this amplitude variation of the sawtooth currents is brought about by mechanical means. Two principal types of arrangements are used. One employs a special type of potentiometer in which the rotation of the antenna is synchronized with the motion of the rotor of the potentiometer. The potentiometer supplies two special outputs, a sine wave and a cosine wave. The other arrangement uses a special type of rotating transformer in which the rotor is synchronized with the antenna.

Sine-Wave Time Base

A basic nonlinear time base that is part of many oscilloscopes is the simple sine-wave sweep. The principal use of sine-wave sweeps is for alignment of tuned circuits. In practically all cases, the frequency of the sine wave is 60 cps and is obtained from the power supply of the oscilloscope. The 60-cps signal is usually procured from the secondary of a filament transformer, as indicated in Fig. 13-28. In this circuit, the 60-cps tap is permanently fixed to the high side of the secondary; when the signal is desired as a time base, it is switched to the horizontal-deflection plates. The 60-cps sine-wave time base is very useful in alignment, because the majority of sweep generators used today employ a 60-cps signal to frequency-modulate the oscillatory signal. A detailed analysis of the use of a 60-cps sine-wave sweep for alignment can be found in Chapter 16.

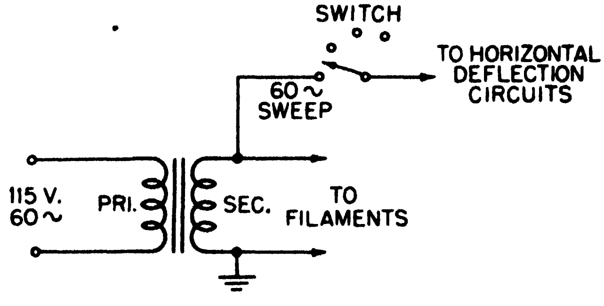


Fig. 13-28.—Circuit for obtaining a simple sine-wave sweep from the power supply of an oscilloscope.

CHAPTER 14

AUXILIARY EQUIPMENT

This chapter covers the use of electronic switching in practical oscilloscope circuits, calibrators, low-frequency time bases, high-voltage supplies, sweep generators, and marker generators. Equipment of this sort is generally termed auxiliary equipment and is used for special applications in radio and electronic work.

ELECTRONIC SWITCHES

An electronic switch opens or closes an electric circuit, having that property in common with ordinary mechanical switches. However, there the similarity ends, for the electronic switch has no moving mechanical parts. Consequently, the frequency of switching can be made much higher than would be possible using mechanical construction; moreover, since there are no moving parts to wear out, maintenance is simplified.

In modern radio and electronic work, an electronic switch may perform many useful functions. Time and space limitations do not permit discussions of all of them, and here we must be limited to some practical circuits and uses.

A practical problem that arises in audio work, for example, is connected with the simultaneous observation of the waveforms of signal voltages in two different stages or circuits. This is shown in Fig. 14-1. The waveform of the input voltage *E1* is observed on oscilloscope No. 1, and that of the output voltage *E2* on oscilloscope No. 2. This is a simple system, but somewhat expensive and inconvenient, since two oscilloscopes are required, and the traces are not close to each other for easy comparison.

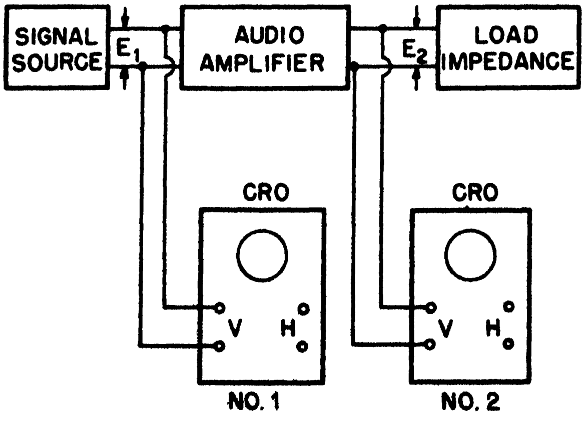


Fig. 14-1.—Simultaneous observation of signal voltages at input and output of an audio amplifier is made possible through this setup.

Using one oscilloscope becomes feasible when a relatively inexpensive electronic switch is employed, as in Fig. 14-2. Here, the electronic switch is shown schematically as a simple single-pole double-throw switch; this is its actual function, although its operation is, of course, a good deal more complex than that. The nature of such a switching system and its electrical makeup will be demonstrated shortly.

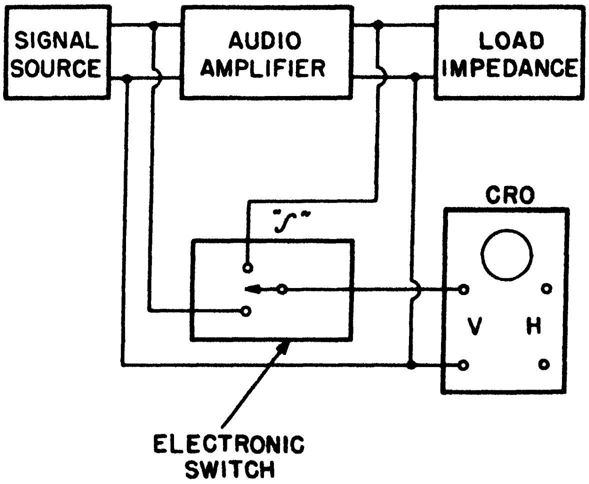


Fig. 14-2.—Compare with Fig. 14-1. Here, a single cathode-ray oscilloscope, in conjunction with an electronic switch, permits simultaneous portrayal of two independent waveforms.

Development of Patterns[1]

Two signal voltages may be shown on the screen of a cathode-ray tube, apparently simultaneously, by means of electronic switching. The circuit principle is shown in Fig. 14-3. The switching rate or frequency should be considerably higher than that of either signal. Double tracing takes place on the screen as shown at (B).

It may be assumed, to illustrate, that when *S* is at *1*, the voltage across the secondary of *T1* produces trace *A* in (B). The length of the trace is determined by the length of time the switch is closed, and its slope by the

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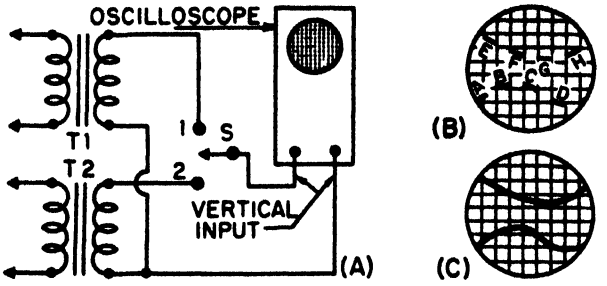
[1] Turner, Rufus P., "Electronic Switch," *Radio News,* vol. 30, p. 28, December, 1943; the following material is an abridged excerpt from this paper.



Courtesy Radio News

Fig. 14-3.—Electronic switch principle is illustrated here.

increment in signal potential. After a short interval, *S* moves to *2* and the secondary voltage of *T2* is applied to the vertical input circuit, producing trace *E*. As the switch alternates from one position to another, a number of separate traces are made, as indicated by *A-B-C-D* and *E-F-G-H* in (B).

The switch movement must be sufficiently rapid to give, through persistence of vision, the illusion of a continuous trace. If the switching frequency is low, the trace will consist of many disconnected lines or dots, as shown in (B). Using a high-frequency switching rate, a pair of apparently continuous line traces as shown in (C) will result.

Two-Channel Switching Circuit

An electronic switching circuit is shown in Fig. 14-4. Triodes *V3* and *V4* are biased somewhat higher than cutoff. A sine-wave voltage applied to terminals *T1* and *T2* is amplified by triode *V1* and is converted into a square-wave voltage by the double-diode clipper tube *V2*. Positive half-cycles of these square waves appear across resistor *R5* and negative half-cycles across *R6*. These half-cycles of square-wave voltage make the grid circuits of *V3* and *V4* alternately conductive, so that if separate signals are applied to terminals *T3-T4* and *T5-T6*, these signal voltages will be delivered alternately by *V3* and *V4* to the vertical plates of an oscilloscope connected to terminals *T7* and *T8*. Signals *1* and *2*, thus, are switched alternately to the oscilloscope at a rate equal to the frequency of the sine-wave voltage applied to terminals *T1* and *T2*.

Complete Circuit

In order to complete the electronic switch circuit for practical application, it is necessary only to add: (1) a power supply, (2) single-stage amplifiers for the two signals, and (3) a single-pole double-throw toggle switch to throw the sine-wave input terminals *T1* and *T2* either to an internal 60-cps source or to terminals for connection to an external sine-wave audio oscillator. The complete circuit, including controls and power supply, is shown in Fig. 14-5.

The sine-wave signal amplifier tube is *V1*. The grid lead of this tube may be switched between terminal *T1* and the "high" side of the 6.3-volt winding of the power transformer. By this arrangement, the switching may be accomplished either at line frequency or at some other desired rate determined by the frequency of an oscillator connected to *T1* and *T2*.

V2 is the 6H6 clipper tube, and *V3* and *V4* are the switching tubes. *V5* is the input amplifier for one observed signal; *V6* the input amplifier for the other. Both of these amplifiers are provided with gain controls (*R7* and *R12*). Each amplifier input is isolated from d.c. by a blocking capacitor.

Signal source No. 1 is connected to terminals *T3* and *T4*, and signal source No. 2 to terminals *T5* and *T6*. *T7* and *T8* are connected respectively to the high and low vertical-input terminals of the oscilloscope.

Du Mont 185-A Electronic Switch and Square-Wave Generator

The circuit diagram of a commercially available two-channel switch is shown in Fig. 14-6. The signal input voltage *E1* may be applied to terminal *A* and ground. A second signal *E2* may be applied to terminal *B* and the common ground circuit. The combined output is

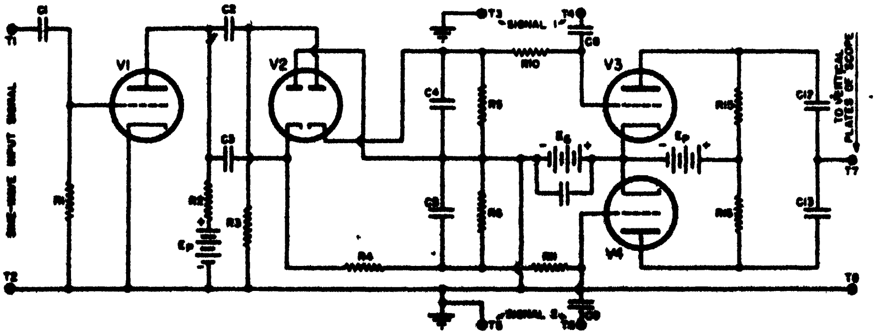


Fig. 14-4.—Electronic switching circuit.

Courtesy Radio News

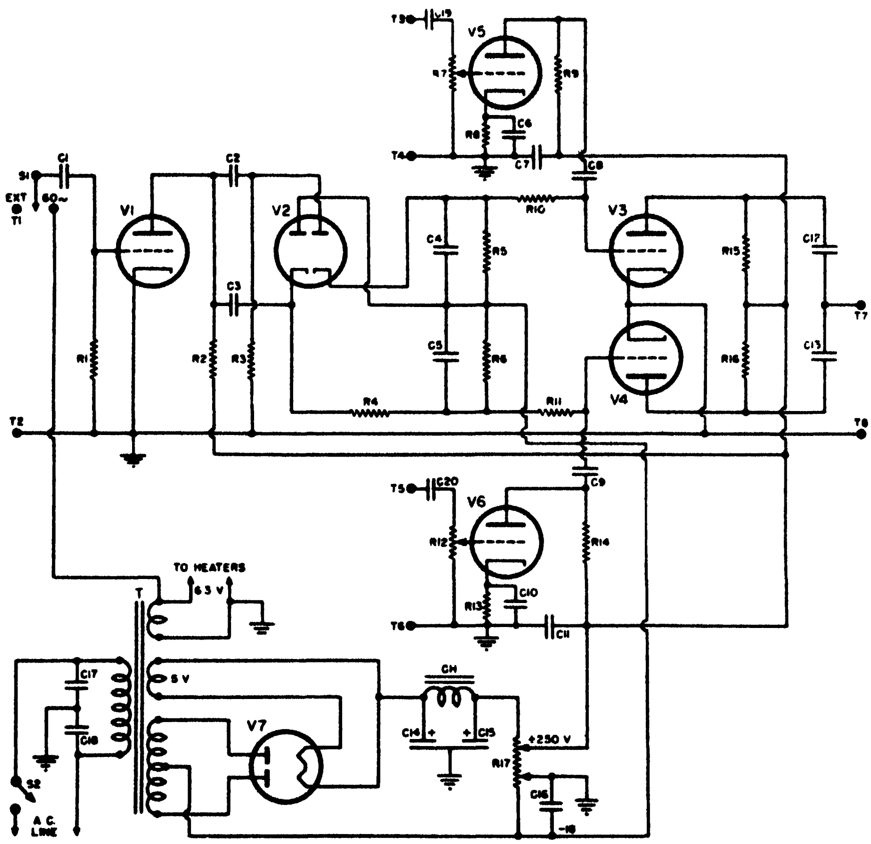


Fig. 14-5.—Complete electronic switching circuit including controls and power supply.

Courtesy Radio News

obtained across *R19-C13* at the output terminal and ground.

As the *E1* signal is applied to *R13* and the *V5* input, the amplitude at the output is a function of the gain of *V5,* which varies according to the operating voltages and the adjustment of *R13.*

It is seen that the cathode of *V5* returns to ground through *R11* and that, therefore, an increase in current through the *V3* cathode and *R11* will increase the bias of *V5,* by making its cathode more positive (equivalent to a more negative grid).

V5 may have its cathode current reduced greatly or cut off by the action of *V3.* To secure a large bias voltage across *R11* and a large *V3* cathode current value, the grid of *V3* must be driven positive. A positive pulse for the *V3* grid is derived from the *V1* plate through

COMPONENT PARTS LIST

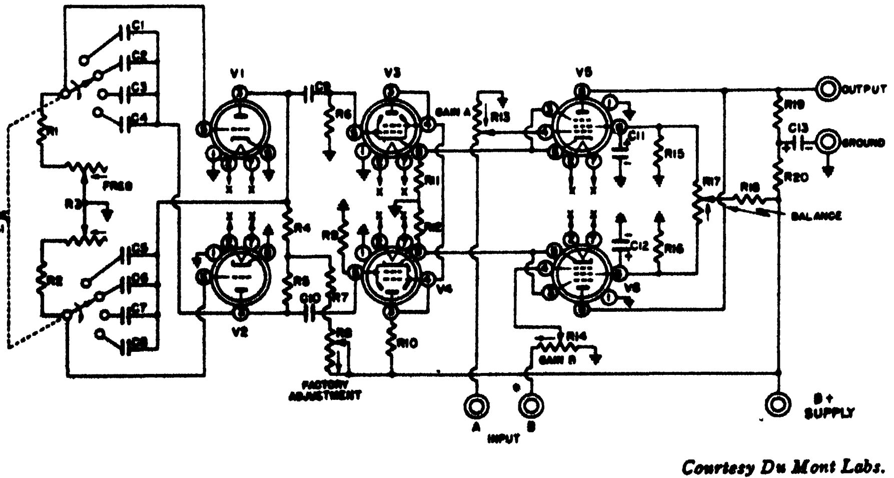
C1—0.01μf. 400V. paper
C2—4700μμf. 500V. mica
C3—0.001μf. 500V. mica
C4—220μμf. 500V. mica
C5—0.01μf. 400V. paper
C6—4700μμf. 500V. mica
C7—0.001μf. 500V. mica
C8—220μμf. 500V. mica
C9—0.5μf. 600V. paper
C10—0.5μf. 600V. paper
C11—4μf. 450V. elec.
C12—4μf. 450V. elec.
C13—24μf. 450V. elec.

R1—470K ½W.
R2—470K ½W.
R3—Dual 4 meg. Pot.
R4—47K 1W.
R5—47K 1W.
R6—1 meg. ½W.
R7—150K 3W

R8—100K pot.
R9—1 meg. ½W.
R10—20K 10W.
R11—2.2K 1W.
R12—2.2K 1W.
R13—100K pot.
R14—100K pot.
R15—100K 1W.
R16—100K 1W.
R17—100K pot.
R18—100K 3W.
R19—47K 3W.
R20—100K 3W.

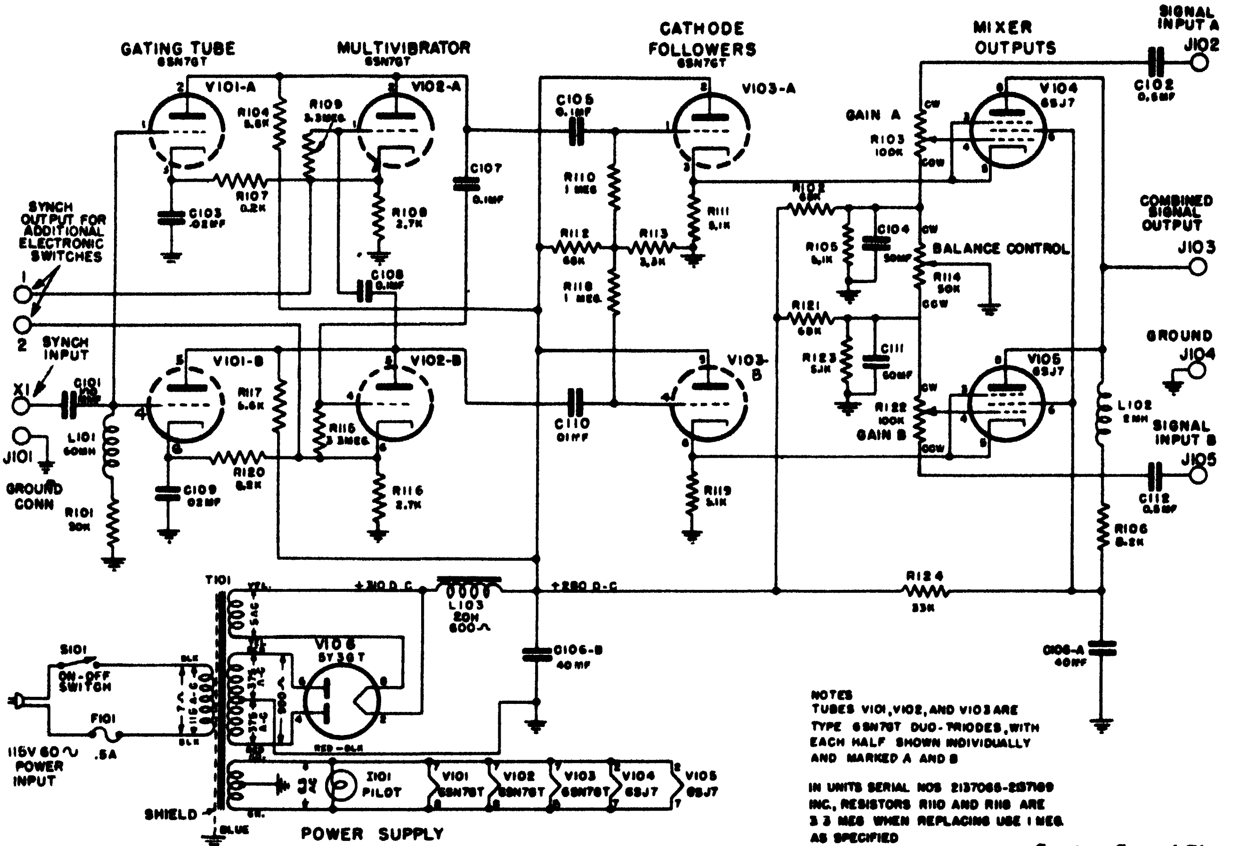
S1—2P 4T. 30° rotary

V1—Type 6J5 or (GT/G)
V2—Type 6J5 or (GT/G)
V3—Type 6V6
V4—Type 6V6
V5—Type 6SJ7
V6—Type 6SJ7



Courtesy Du Mont Labs.

Fig. 14-6.—Du Mont 185-A electronic switch and square-wave generator.



NOTES
TUBES V101, V102, AND V103 ARE TYPE 6SN7GT DUO-TRIODES, WITH EACH HALF SHOWN INDIVIDUALLY AND MARKED A AND B

IN UNITS SERIAL NOS 2137066-2137169 INC., RESISTORS R110 AND R118 ARE 3.3 MEG WHEN REPLACING USE 1 MEG AS SPECIFIED

Courtesy General Electric

Fig. 14-7.—General Electric YE-9 electronic switch.

C9. This pulse occurs during one-half of the signal cycle of operation of *V1*.

V1 and *V2* together form a simple, standard, variable-frequency multivibrator system. The frequency variations are controllable by means of the variable resistances marked *R3* and the capacitance selection system.

V2 serves as a signal source for *V4*, in a manner similar to that described for the *V1-V3* combination, but opposite in phase. Similarly, *V4* serves *V6* in a manner identical to that of *V3* with respect to *V5*.

V5 and *V6* thus serve as signal *gates* or switches, controlled by the multivibrator pulse signals generated in the *V1-V2* stage. These gating tubes then pass or block the signals in accordance with the controlling impulses furnished by the pulsing stage.

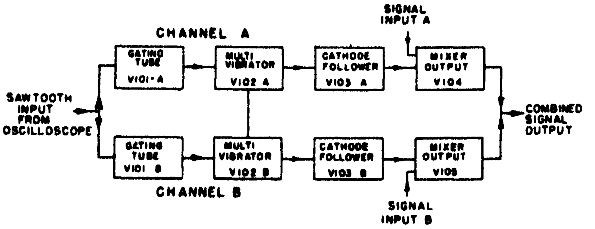
General Electric YE-9 Electronic Switch

Another commercially made electronic switch is the General Electric YE-9. This system operates, so far as the gate tubes are concerned, in a manner similar to the one previously described. The complete circuit is shown in Fig. 14-7. *V104* and *V105* are the signal gates. (*V101*, marked "gating tube," gates the synchronizing signal to the proper half of *V102*.)

The signal input at terminal *J105* feeds to *R122* through *C112*. That at *J102* feeds through *C102* to *R103*. *R122* and *R103* permit adjustment of the individual signal outputs which are combined at *J103*.

V103-A controls *V104* and receives its grid excitation through *C105* from the *V102-A* multivibrator plate. *V103-B* controls *V105* and is voltage-driven through *C110* by *V102-B*. The synchronizing input voltage appears across *X1* and *J101* and is coupled to the *V101-A* and *V101-B* grids through *C101*.

A simplified block diagram of the equipment is shown in Fig. 14-8. In effect, this circuit consists of two channels — channel *A* and channel *B*. The two channels are identical but operate alternately, and at half the frequency of the sawtooth input from the oscilloscope. Each channel generates a square wave, which is out of phase with that generated in the other channel. These in turn are used to *gate* (control the off-on operation of) the individual mixer output tubes, so that the individual signal inputs are alternately delivered to the output terminal. Due to the persistence of vision of the human eye, and the persistence characteristic of the fluorescent material of the oscilloscope screen, the two signals appear on the screen simultaneously.



Courtesy General Electric

Fig. 14-8.—Simplified block diagram of General Electric YE-9 electronic switch.

Frequency-Dividing Multivibrator

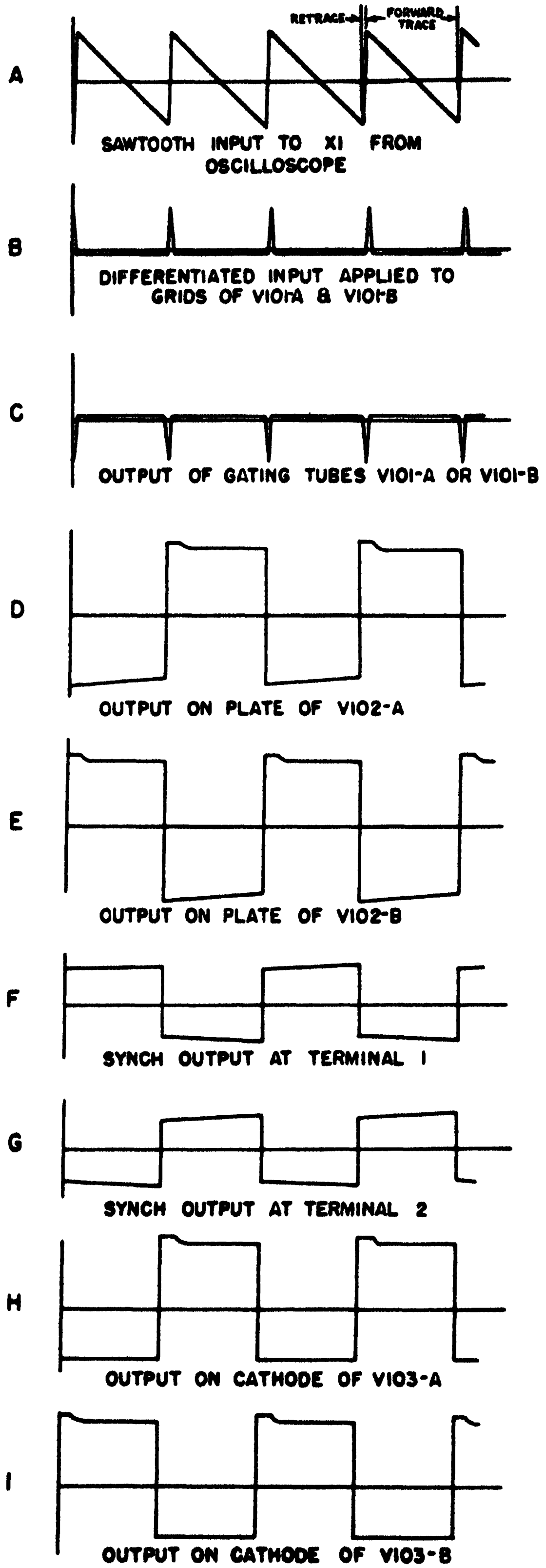
For the sake of circuit analysis, each half of the 6SN7GT tubes (*V101, V102,* and *V103*) will be considered individually. For example *V101-A* will be considered as one triode tube, and *V101-B* as another.

The frequency-dividing multivibrator consists of gating tube *V101* and multivibrator tube *V102,* and their associated components. The function of this circuit is to produce two 50 per cent square waves, 180° out of phase with each other, so that each has one-half the repetition frequency of the sawtooth input from the oscilloscope.

In use, the sawtooth input from the oscilloscope is received at terminal *X1.* This sawtooth input has an appearance as illustrated in Fig. 14-9A. A differentiation circuit, composed of *C101, L101,* and *R101,* differentiates the sawtooth input into a waveform illustrated in Fig. 14-9B.

Disregarding tubes *V101-A* and *V101-B* for the moment, *V102-A* and *V102-B* form a conventional back-coupled multivibrator having a free-running frequency of approximately one cps. Therefore, at a given time (neglecting the switching time of the multivibrator), either tube *V102-A* or *V102-B* will be conducting, while the other will be cut off. For purposes of explanation, let tube *V102-A* be conducting, and *V102-B* be cut off.

The plate current of *V102-A* produces a voltage drop across cathode resistor *R108* of approximately 30 volts; this voltage is transferred through *R107* to the cathode of gating tube *V101-A,* and cuts off *V101-A.* Since no current is flowing through *V102-B,* tube *V101-B* will not be cut off. The positive pulses (Fig. 14-9B), derived by differentiating the sawtooth input, are applied directly to the grids of *V101-A* and *V101-B.* However, at the particular instant of time depicted, only *V101-B* is conducting and will amplify


RETRACE
FORWARD TRACE
A
SAWTOOTH INPUT TO X1 FROM OSCILLOSCOPE
B
DIFFERENTIATED INPUT APPLIED TO GRIDS OF V101-A & V101-B
C
OUTPUT OF GATING TUBES V101-A OR V101-B
D
OUTPUT ON PLATE OF V102-A
E
OUTPUT ON PLATE OF V102-B
F
SYNCH OUTPUT AT TERMINAL 1
G
SYNCH OUTPUT AT TERMINAL 2
H
OUTPUT ON CATHODE OF V103-A
I
OUTPUT ON CATHODE OF V103-B

Courtesy General Electric

Fig. 14-9.—Waveforms in General Electric YE-9 electronic switch system.

the pulse. The output of *V101-B* is a negative pulse as illustrated in Fig. 14-9C (inverse of Fig. 14-9B), which appears across the load resistor *R117*. This negative pulse is transmitted to the grid of *V102-A*, cutting off its plate current, and, by multivibrator action, turning on *V102-B*. Thus, the first positive pulse (of Fig. 14-9B) causes the multivibrator to switch once. Similarly, the second positive pulse will affect *V101-A*, which will apply a negative pulse to *V102-B*, cutting it off, which will turn on *V102-A*. Thus the circuit has been restored to its original state and one cycle has been completed. Since one cycle requires two input pulses (differentiated oscilloscope sawtooth input), it will produce an output wave having one-half the frequency of the oscilloscope sweep frequency.

Integrating networks *R107* and *C103*, and *R120* and *C109*, effect a slight delay of the gate voltage applied by the multivibrator to the gating tubes, which gives the multivibrator time to switch. Capacitors *C103* and *C109* also prevent degeneration to pulse amplification in tubes *V101-A* and *V101-B*. The square-wave outputs of multivibrator tubes *V102-A* and *V102-B* are illustrated in Fig. 14-9D and E. The two outputs are out of phase with each other and their frequency is equal to one-half the frequency of the sawtooth input.

Square-wave outputs are taken from the cathodes of *V102-A* and *V102-B* and may be used to synchronize additional electronic switches. These outputs are illustrated in Fig. 14-9F and G, and may be taken from terminals *1* and *2*.

Cathode Followers

Each half of the type 6SN7GT duo-triode *V103* is employed as a conventional cathode-follower tube. *V103-A* serves channel *A*, and *V103-B* serves channel *B*. The function and operation of each cathode follower is the same except that they operate alternately. Each cathode follower serves three purposes:

a. It squares the negative half-cycles of the multivibrator output waves shown in Fig. 14-9D and E, prior to applying them to the mixer tube.

b. It provides a low-impedance source for driving the cathodes of the mixer.

c. It serves to isolate signal variations in the mixer from the multivibrator.

Squaring is accomplished by driving the grids of the cathode followers beyond cutoff with the negative half-cycles of the square waves illustrated in Fig. 14-9D and E. The square-wave outputs of the cathode followers are illustrated in Fig. 14-9H and I. Notice that the output of *V103-A* is in phase with the output of *V102-A*, and the output of *V103-B* is in phase with the output of *V102-B*. Also notice that the outputs of *V103-A* and *V103-B* are out of phase with each other.

Mixer Outputs

There are two mixer output tubes *V104* and *V105*—one for channel *A* and one for channel *B*. The function of each mixer is to apply the signal received at its grid, alternately, to the output terminal.

As shown on the schematic diagram, Fig. 14-7, signal input *A* is applied to the control grid of mixer tube *V104* through the "Gain A Control"; and signal input *B* is applied to the control grid of mixer tube *V105* through the "Gain B Control." The outputs of the two tubes are connected together to the combined signal-output terminal *J103*. Alternate application of the applied signal inputs to the combined signal-output terminals is accomplished by gating the cathodes of the mixer tubes with the square-wave outputs of the cathode-follower tubes. Since the square-wave outputs of the cathode-follower tubes (Fig. 14-9H and I) are 180° out of phase with each other, one mixer tube will be conducting while the other will be off. For this reason, only one signal appears at the output terminal at a time.

Notice that the output terminal *J103* is connected directly to the plates of the mixer output tubes, and is at 130 to 160 volts d-c potential, depending upon the position of the "Balance Control."

The "Balance Control" is provided to permit vertical separation or superposition of the two traces on the oscilloscope. Balance is obtained by varying the quiescent (no-signal) grid bias, and consequently the quiescent plate current of the mixer tubes. For superposition of the two traces, the "Balance Control" is adjusted to give equal plate currents through the two mixer tubes. In order that adjustment of the "Balance Control" may not be critical for low values of signal input, the control is effectively tapered by the addition of the two resistors *R105* and *R123*.

To minimize nonlinear distortion, negative feedback is introduced in the mixer tubes by the omission of bypass capacitors across the cathode resistors *R111* and *R119*. The feedback is approximately 62 per cent at low and medium frequencies. Inductance *L102* in the plate lead provides video compensation at high frequencies.

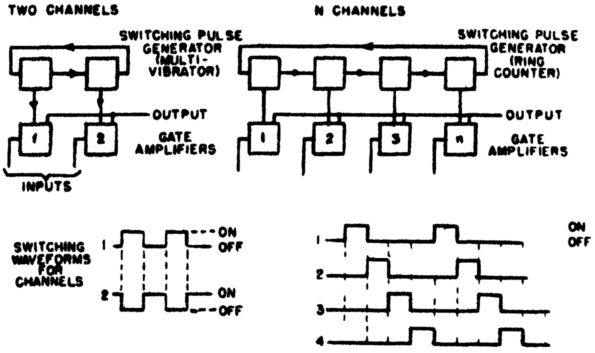
Power Supply

The power-supply circuit is shown in the lower portion of the schematic diagram, Fig. 14-7. It comprises a type 5Y3GT rectifier tube (*V106*) in a conventional full-wave rectifier circuit. A single-section inductance-input filter composed of *L103* and *C106-B* provides a d-c output of 280 volts for the plates of *V101, V102,* and *V103*. Additional filtering is provided by *R124* and *C106-A,* for the plates and screens of the mixer output tubes *V104* and *V105*. Sufficient bleeder current flows through resistors *R112, R102,* and *R121* (68,000 ohms each) to prevent the output voltage of the power supply from rising to excessive values after the closing of the "On-Off" switch, during the warm-up time of the tubes.

A 6.3-volt secondary winding of power transformer *T101* supplies all filament power, and also excites pilot lamp *I101*.

Four-Channel Electronic Switch[2]

Although the two-channel switch just described is adequate for many applications, there are occasions when a greater number of channels must be handled. Fig. 14-10 shows a block diagram of a four-channel electronic switch. The *ring-counter* circuit can be developed for *n* channels as indicated in the drawing.


TWO CHANNELS
N CHANNELS
SWITCHING PULSE GENERATOR (MULTI-VIBRATOR)
SWITCHING PULSE GENERATOR (RING COUNTER)
OUTPUT
OUTPUT
1
2
GATE AMPLIFIERS
1
2
3
n
GATE AMPLIFIERS
INPUTS
SWITCHING WAVEFORMS FOR CHANNELS
1
2
ON
OFF
ON
OFF
1
2
3
4
ON
OFF

Courtesy Electronics

Fig. 14-10.—Block diagram of a four-channel electronic switch.

In an ordinary two-channel electronic switch, a multivibrator supplies the required square-wave switching pulses, which alternately bias to cutoff the respective gate amplifiers. The common plate connection between these amplifiers provides the point at which the composite signal is applied to the oscilloscope. On the other hand, the *n*-channel electronic switch contains *n* generators which supply switching pulses consecutively to gate amplifiers 1 to *n*, this cycle of operation being continuously repeated. The duration of each switching pulse must obviously be 1/*n* times the period for each cycle of operation.

Ring Counter

The requirements for a switching pulse generator have been conveniently satisfied by the use of an electronic ring counter. By means of this system, it becomes theoretically possible to devise an electronic switch for any number of channels whatsoever. Practical considerations restrict the application to four channels.

An electronic ring counter is essentially a group of square-wave generating circuits so arranged that only one circuit of the group is "on" at any instant. Trigger pulses are applied simultaneously to all the circuits. At each pulse, the next succeeding member of the group is turned on, and the one previously operating is turned off. Effectively, continually repeated trigger pulses cause the "on" conditions to move around the chain or ring of square-wave generators; therefore, it is called a *ring counter*.

This cycle of operation is continuously repeated. Since all gate amplifiers are normally biased beyond cutoff, the function of these positive switching pulses is to turn on the respective gate amplifiers. Any one of the gate circuits is in the conducting state for a period equal to 1/*n* times the switching cycle, and is off during the remaining part of the cycle.

The phase relations between the switching pulses for a four-channel electronic switch are shown in Fig. 14-10. The composite waveform is again derived from the common plate connection between gate amplifiers. Oscillograms of the waveforms actually obtained are reproduced in Fig. 14-11.

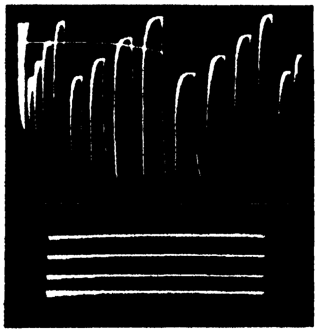


Fig. 14-11.—Oscillograms of waveforms actually obtained from a four-channel electronic switch.

Courtesy Electronics

It may seem that considerable resolution of the waveforms under study will be lost because of the relatively short period during which any channel is active. This would be important if the switching rate were comparable to, or less than, the frequencies contained in the waveforms under study. By making the switch-

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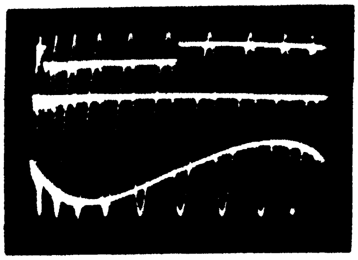
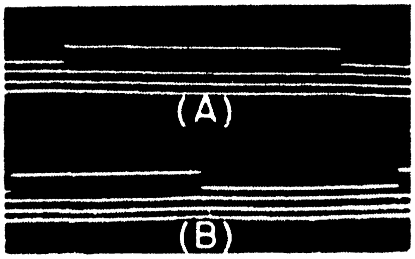
[2] Moerman, N. A., "Four channel electronic switch," *Electronics*, vol. 19, p. 150, April, 1946. The following material is an abridged excerpt from this paper.



Courtesy Electronics

Fig. 14-12.—Oscilloscope traces resulting when three different signals were introduced through the four-channel switch. Top trace presents a 60-cps square wave and the bottom a 60-cps sine wave. No signal was applied to the second trace input. A 600-cps sine wave of large amplitude superimposes the third trace upon the others.



Courtesy Electronics

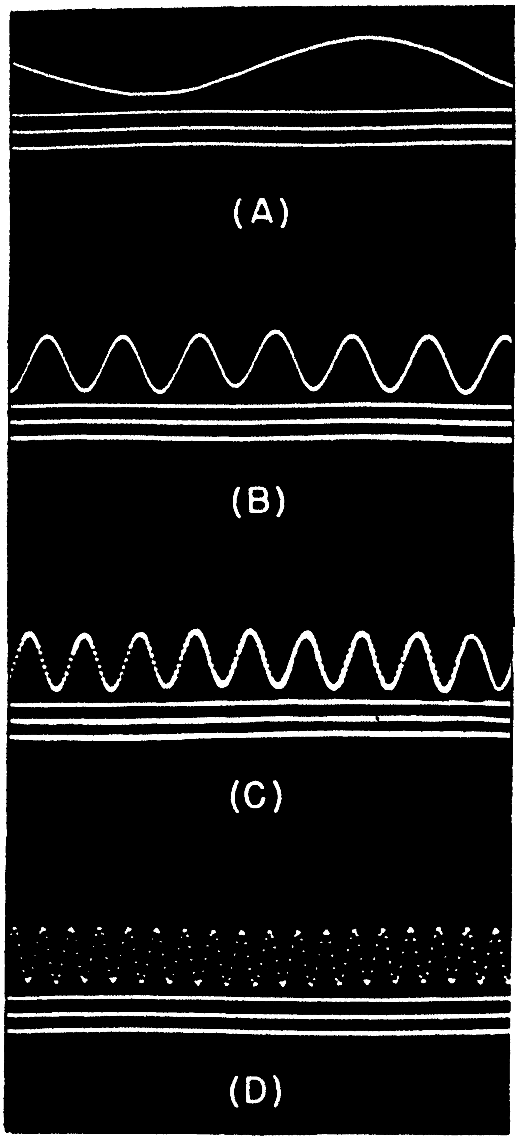
Fig. 14-13.—60-cps square-wave recordings taken with a revolving-drum type camera. In (A) the camera speed is 285 inches per second and in (B) the speed is 195 inches per second.

ing rate as high as possible, this difficulty is minimized; a switching rate of 25,000 cps, for example, is a practical figure. The degree of resolution actually obtained with the four-channel electronic switch is shown in Figs. 14-12, 14-13, and 14-14.

Design

The chassis contains all the essential elements for the operation of the four-channel electronic switch with the exception of power supplies and the necessary preamplifiers. These elements are: (1) the trigger pulse source; (2) the switching-pulse generator (ring counter); (3) the gate amplifiers; and (4) the output amplifier.

For proper operation of the ring counter, a source of trigger pulses of the proper frequency and shape is required. A pulse frequency of 100 kc, which yields a switching rate of 25,000 cps is used. In the complete four-channel switch circuit in Fig. 14-15, one half of tube *V2* (6SN7) is connected with associated components to form a Pierce oscillator. The other half of *V2* serves as a square-wave amplifier.



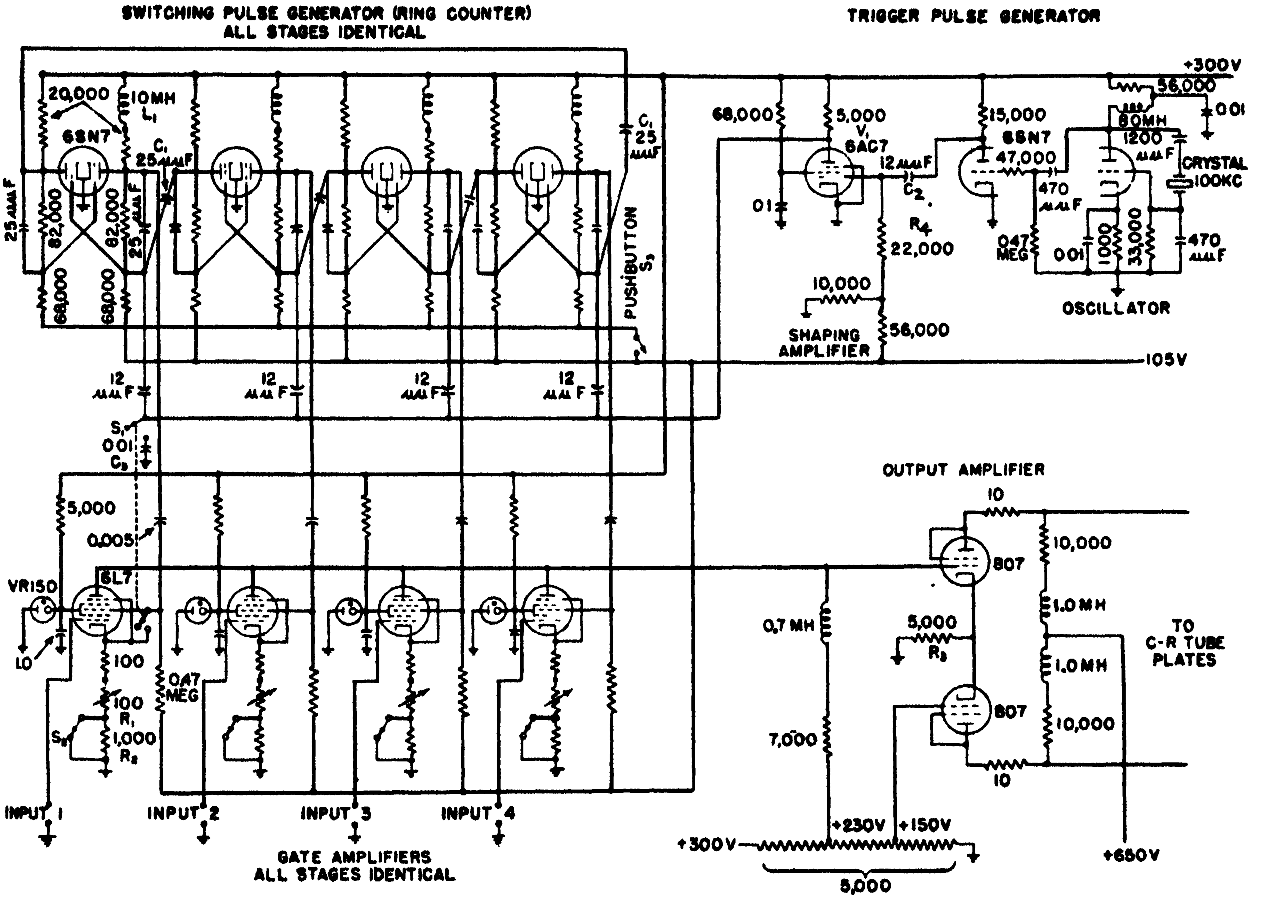
Courtesy Electronics

Fig. 14-14.—Photographic recordings of signals fed to a commercial oscilloscope through the four-channel switch, only one channel of which is used in each case: (A) 100 cps; (B) 500 cps; (C) 1,000 cps; and (D) 2,000 cps.

Circuit Details

Differentiation of the 100-kc square waves is accomplished by the *R-C*-combination composed of *R4* and *C2*. Of the positive and negative pulses resulting, only the positive pulses are amplified by tube *V1*, which is biased to cutoff. Negative pulses appearing at the plate of *V1* have the desired characteristics. These are applied to a common bus for injection into the ring counter and gate amplifier.

The switching pulse generator (ring counter) is composed of four 6SN7 tubes connected in standard



Courtesy Electronics

Fig. 14-15.—Complete four-channel switch circuit.

Eccles-Jordan square-wave circuits with the addition of inductance *L1* in each plate circuit. The inductance is placed in that side of the circuit from which the positive switching pulse is derived and its use permits a considerable improvement in the squareness of the switching pulse results without adversely affecting normal operation of the circuits.

Generator Circuit

An Eccles-Jordan circuit (sometimes called a flip-flop circuit) has two stable states such that, for either state, one grid is sufficiently biased to cut off tube current, and the other grid is at zero, permitting current flow. Of those grids in the ring counter that are connected to the source of trigger pulses, only one grid is zero at any instant. This grid is associated with the circuit that is "on."

The trigger pulse operates on this grid and flops the circuit to its other stable state, "off." During this transition, the positive rise in voltage at the left-hand plate is applied through *C1* to the cutoff-biased grid of the succeeding circuit, which is then turned on. The next trigger pulse operates on this circuit, which is then flopped to the off position, as in the case of the preceding circuit.

In this way, a positive switching pulse is generated at the right-hand plate of each circuit during the sequence off-on-off. It will be noted in the schematic diagram that pushbutton switch *S3* has been inserted between the two legs of the square-wave circuits, one of which is connected in reverse order. This provides a front-panel control for initiating the proper sequence of operation for the counting ring after applying power.

Switching the Pulses

The gate amplifiers serve the twofold purpose of switching and then mixing the signals applied to their respective inputs. For this function, the type 6L7 tube was selected. At the outset, it was felt that the inner control grid of the 6L7 would be the logical point for introducing the switching pulse, since cathode-current cutoff could then be assured. This arrangement was tried and found to be unworkable because of the distortion introduced by applying the useful signal to the second control grid. The functions of the two control grids were then reversed and a great improvement in the operation noted. The disadvantage to this mode of operation, however, is the incomplete cathode-current cutoff, requiring isolation of the screen-voltage supplies to insure minimum couplings.

Variable resistor *R1* in the cathode circuit provides a means for positioning the trace on the cathode-ray tube with respect to the other traces. When it is desired to observe less than the four available traces, switch *S2* is opened. This introduces a large resistor *R2* into the cathode circuit, which then deflects the trace from the corresponding tube off the screen.

For added flexibility, provisions are made for restoring normal single-trace operation. This may be desired where high-frequency recording and good resolution are required, and is made possible by operating switch *S1*. This shunts the trigger pulses to ground through *C3* and removes the bias on the operating 6L7 tube by connecting its second control grid to the cathode.

Wideband Amplifier

It was evident early in the development of the apparatus that amplification of the composite signal would present some difficulty. The problem involved is that of amplifying unidirectional square-wave pulses of approximately 75 volts amplitude. These pulses are modulated by as much as 25 per cent. Capacitance coupling could not be used because of the transient d-c amplification requirement. Furthermore, it was found to introduce interchannel coupling.

The output amplifier circuit finally evolved is shown in the diagram, Fig. 14-15. Triode-connected 807 tubes in push-pull were selected because of their large signal-handling capabilities, coupled with low distortion. An excessively large common cathode resistor *R3* is employed which serves two important functions: The cathode potential substantially follows the grid, thereby absorbing the large d-c potential at the 6L7 plates and minimizing undesirable loading due to the direct connection. Phase inversion is also conveniently provided.

The grid voltage of the phase inverter tube is selected at a point on a voltage divider from B+ to ground, to permit proper placement of the composite signal on the cathode-ray-tube screen. There is linear response over the useful portion (about four-fifths) of the screen. Contributing to this is the high value of plate voltage (650 volts) applied to the 807 amplifiers.

Recording

For recording, use is made of a Du Mont type 5LP5 (blue screen) cathode-ray tube, operating at maximum voltage in order to attain maximum spot intensity. A revolving-drum type camera, having an *f*2.3 lens, of three-inch focal length, records a one-to-one image on a single strip of Eastman No. 697 oscillograph recording paper.

Records taken with this apparatus are illustrated in Figs. 14-13 and 14-14. The cathode-ray-tube voltages were, in this case, only three-quarters of the maximum value. These records demonstrate the degree of resolution and recording intensity that may be obtained.

Interpretation of Records

It may appear disadvantageous that the steeply sloping portions of recorded transients are defined by a series of dots. This characteristic, however, may be used to advantage where it is desired to obtain quantitative values for these slopes. For this purpose, the oscillator is crystal-controlled, thereby accurately fixing the time between successive dots. The slope measurement then becomes that of simply measuring the vertical distance between successive dots and dividing by 1/25,000 second. There is an increase in precision attainable in this way as compared to the normal method, in which the horizontal component of steep slopes must be determined by linear measurement.

OTHER OSCILLOSCOPE ACCESSORIES

Since the oscilloscope is such a versatile instrument, it is natural that numerous accessories should have been developed for it, for particular requirements. The several types of electronic switches just discussed are very useful, but there are others which must be considered here.

Du Mont Voltage Calibrator, Type 264-A

Since no attempt is ordinarily made, in oscilloscope design, to calibrate the spot deflections as a function of

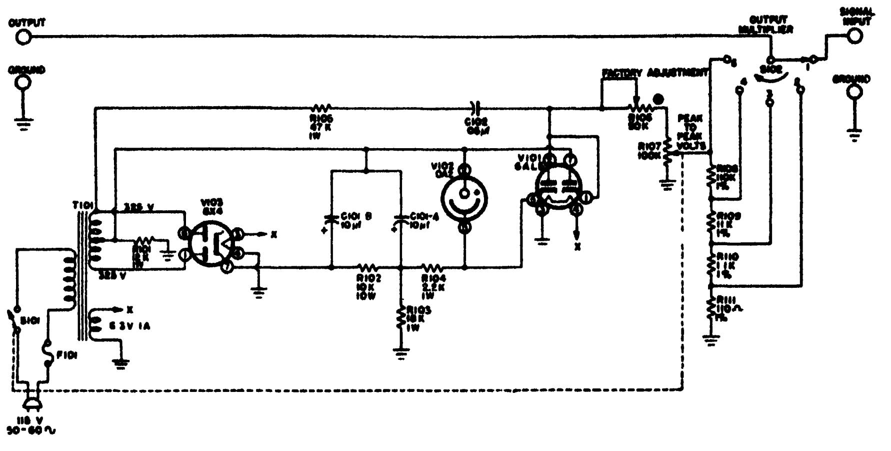


Fig. 14-16.—Circuit of the Du Mont voltage calibrator, Type 264-A.

Courtesy Du Mont Labs

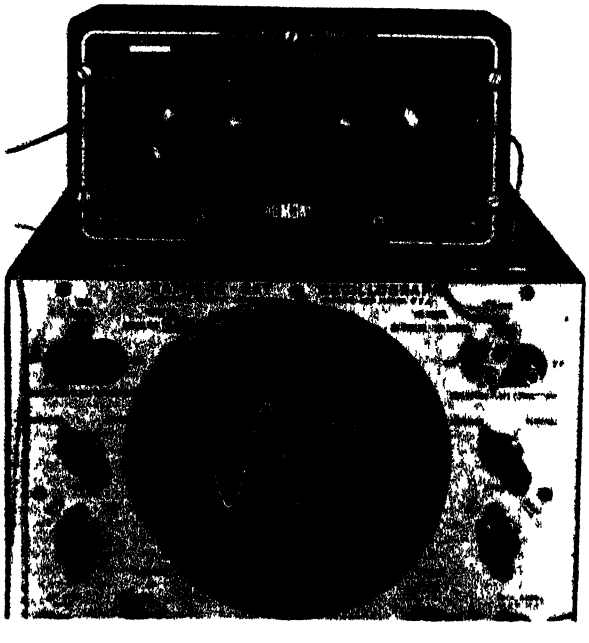
voltage, chiefly because of the manual amplitude adjustment and normal amplifier and tube variations, some means of determining the voltage for a given deflection amplitude is often desired. The Du Mont Voltage Calibrator, Type 264-A, is such a device. Its circuit is shown in Fig. 14-16.

The circuit consists of a power supply, operated from a 115-volt 50-60-cps line voltage, which furnishes equal positive and negative bias voltages to a dual-diode. One of these diodes is connected to clip the positive half-cycle of the sine-wave output of the power transformer, and the other diode clips the negative half-cycle at the proper voltage to result in exactly 100 volts peak-to-peak at the output terminals. A miniature voltage-regulator tube maintains the proper bias on the two diodes. The output voltage is a clipped sine wave with essentially flat tops.

The power supply consists of a full-wave rectifier, *V103*, and an *R-C* capacitor-input filter consisting of *C101A*, *C101B*, *R102*, *R103*, and *R104*, with voltage regulator *V102* connected across the output. A negative voltage is picked off a resistor which is inserted between the center tap of the power transformer and ground. The output of the power supply is taken between this negative voltage and the positive voltage, and the two filter capacitors *C101A* and *C101B* serve to filter both voltages. Although the difference in potential between the two output leads of the power supply is 150 volts, one of these leads is +75 volts with respect to ground and the other is −75 volts with respect to ground.

The dual diode, with its two sections oppositely biased, clips each half-cycle of the output voltage at 75 volts, resulting in an output voltage of 150 volts peak to peak.

Then, the output voltage is adjusted to a maximum of 100 volts at the factory by adjustment of a variable resistor *R106*. The full output voltage is obtained at position *5* of the output amplifier switch *S102*, with the tap on *R107* being at the end of the resistor farthest from ground. Attenuation of the output voltage is accomplished with the continuous output attenuator *R107* and the four steps of the decade attenuator obtained with resistors *R108*, *R109*, *R110*, and *R111*. An additional point on the attenuator switch is provided to connect the signal under observation directly to the oscilloscope, and disconnect the calibrator output; this makes a change of interconnection or an external switch unnecessary. The normal procedure is to feed the signal under study to the input terminals of the calibrator. The output terminals are then con-



Courtesy Du Mont Labs.

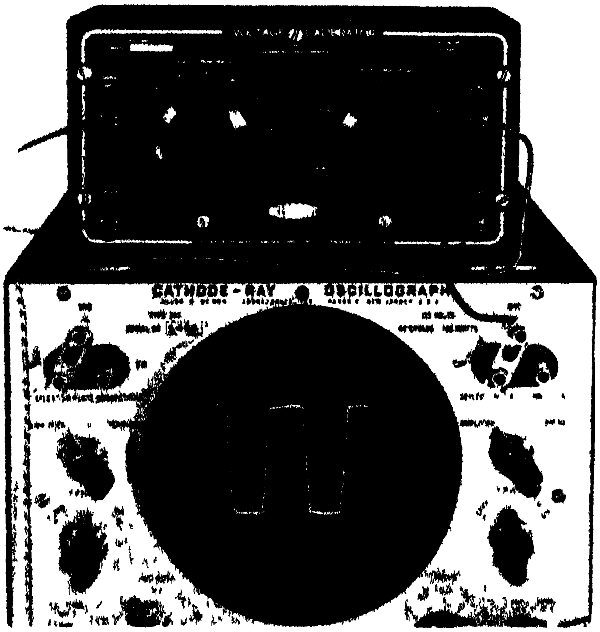
Fig. 14-17.—Du Mont Type 264-A voltage calibrator in use with oscilloscope, with the "Output Multiplier Switch" at "Signal Output" position.

nected to the input of the oscilloscope. By operating the attenuator switch, either the unknown signal or any one of four ranges of calibrating voltage may be applied to the input of the oscilloscope.

To measure an unknown signal, turn the "Output Multiplier" switch to the "Signal Output" position and adjust the gain control of the oscilloscope to produce a pattern of convenient size on the screen of the cathode-ray tube. Note the peak-to-peak deflection on the screen; the cross-ruled scale is useful for this.

Turn the "Output Multiplier" switch to the proper range. (The proper range is that which enables the "Peak-to-Peak Volts" to be set as near the high end as possible.) Then, without changing the setting of the vertical gain control of the oscilloscope, adjust the calibrated output dial "Peak-to-Peak Volts," to result in the same peak-to-peak amplitude of the calibrating voltage as that produced by the signal When this adjustment has been carefully made, read the calibrated dial and multiply by the decimal multiplier to obtain the peak-to-peak signal voltage.

Figs. 14-17 and 14-18 show the Type 264-A Voltage Calibrator in use with a cathode-ray oscilloscope. In Fig. 14-17, the signal voltage appears on the screen of the cathode-ray tube. Note that the "Output Multiplier" switch is at the "Signal Output" position. In Fig. 14-18, the switch has been turned to the ×*.1* position and the calibrating voltage appears on the screen. The amplitude of the calibrating voltage has been set to be the same as that of the signal voltage, and appears on the dial as 18.5 volts. The actual peak-to-peak voltage is, then, 18.5 × 0.1 = 1.85 volts.



Courtesy Du Mont Labs.

Fig. 14-18.—Same as Fig. 14-17, but with switch turned to ×*.1* position.

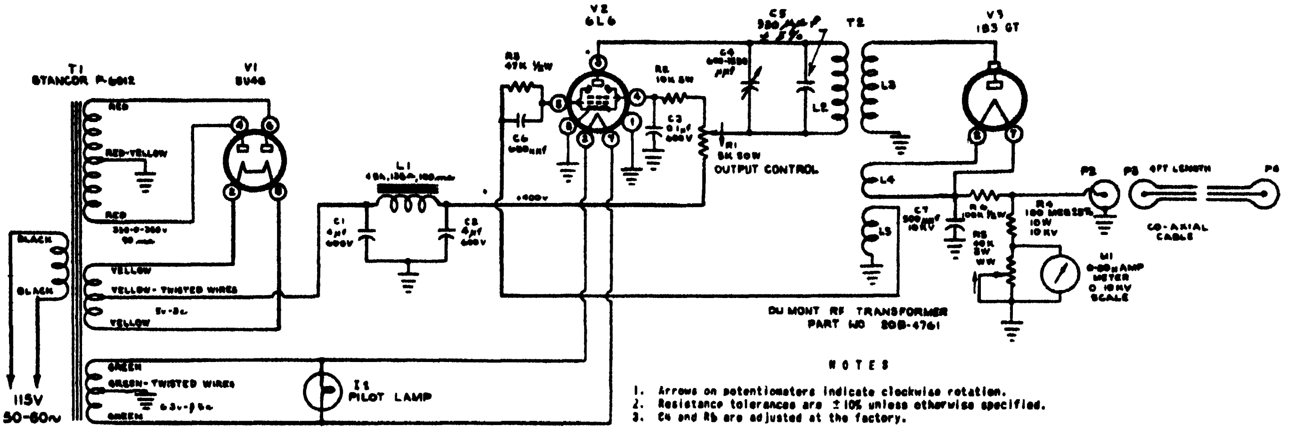
Du Mont High-Voltage Supply, Type 263-A

This supply has been designed primarily as an accessory for cathode-ray oscilloscopes, extending their usefulness by providing higher accelerating potentials. The application of a high accelerating potential to the cathode-ray tube produces an intense pattern which is very desirable for photographic or projection purposes.

Unfortunately, the intensifier voltage of standard cathode-ray tubes cannot be increased much above twice the second-anode voltage, without producing serious pattern distortion. This difficulty has been overcome by the development of the Type 5RP cathode-ray tube, which allows ratios as high as 10:1 without distortion. The Type 263-A High-Voltage Power Supply may be employed to supply intensifier voltage in any application where a Type 5RP is used; it may be used, of course, in other cases where a portable, rugged, and safe high-voltage supply of its rating is needed. It provides, by rectifying a high radio-frequency voltage, a d-c voltage output which is continuously variable from 5,000 to 10,000 volts with loads up to 200 microamperes.

The instrument consists of a medium-voltage rectifier (600 volts), a radio-frequency oscillator, an r-f step-up transformer, a half-wave rectifier, and a high-voltage filtering and metering system. Upon the front panel is mounted an output kilovoltmeter reading up to 10 kv, an output connector, a control for varying the voltage, two ground terminals, pilot light, fuse, and power switch. A four-foot length of shielded output cable and connector is furnished with the unit. The circuit diagram is given in Fig. 14-19. The principles of operation are quite familiar, but there are some new applications of these principles which are rather interesting. Ordinary power supplies have a step-up transformer for increasing the power-line voltage, a rectifier, and filter for producing suitable d-c output voltage. The Type 263-A High-Voltage Power Supply is similar to this, with the exception of operating frequency; instead of a power source frequency of 60 cps, it operates on 200 kc; it is, moreover, a half-wave rectifier, rather than the customary full-wave.

The 200-kc input to the rectifier is obtained from a 6L6 oscillator of the "tickler" type, with a 400-volt plate supply of usual form. The load is taken from the plate coil of the oscillator by winding *L3* of trans-



Courtesy Du Mont Labs.

Fig. 14-19.—Circuit of Du Mont high-voltage supply, Type 263-A.

formers *T2*, *L2* being the plate coil and *L5* the tickler feedback. The load secondary *L2* is a high-*Q* coil with a natural resonant frequency near that of the oscillator. This gives a high-voltage step-up.

Voltage is conveniently controlled by varying the plate voltage of the series-fed oscillator by means of potentiometer *R1*. For the maximum output of 10 kv, the oscillator operates with maximum plate voltage and generates approximately 250 volts rms at 200 kc across the tank coil. The corresponding secondary voltage of approximately 7,000 volts rms is equivalent to 10 kv peak volts.

This voltage is rectified by a 1B3-GT, which functions as a half-wave rectifier. Filament voltage for this rectifier tube is obtained from a two-turn coil *L4*, which is coupled to the high-voltage secondary of the r-f transformer. Although the filament voltage varies as the output voltage is changed, the regulation and current rating of the supply remains sufficiently good over the range from 5 to 10 kv. Adequate filtering is provided by the filter capacitor *C7*, which is connected across the output and reduces the ripple voltage to less than 0.5 per cent. The capacitance of the output cable (200 $\mu\mu f$) serves as an additional filter, and further reduces the ripple voltage.

Output voltage is indicated by microammeter *M1* in the bleeder circuit. This meter and its calibrating shunt *R5*, in series with *R4*, form the bleeder. The meter is calibrated in kilovolts so that its deflection, which is proportional to the bleeder current, is a direct indication of the output voltage.

Du Mont Linear Time-Base Generator, Type 215

Oscilloscopes, generally, have built-in linear sweep generators, which cause the spot to traverse the screen at a uniform horizontal rate, returning very rapidly to repeat the sweep. If the sweep frequency is reduced, the linearity may become poor; moreover, the sweep rate can not generally be reduced below, perhaps, four or five per second.

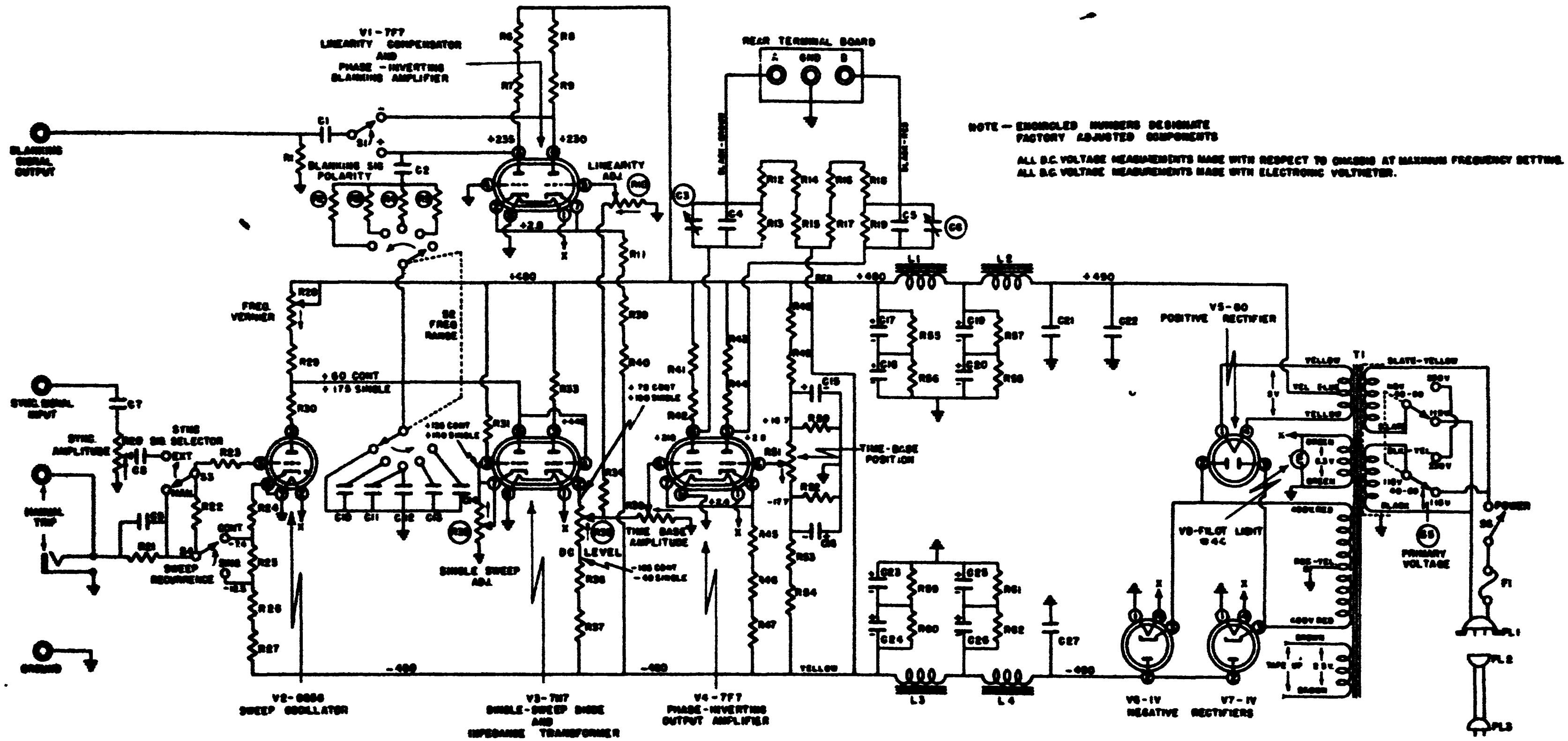
An instrument, whose circuit is given in Fig. 14-20, has been designed to go beyond these limits, to apply a sweep voltage of accurate linearity over the operating range of approximately 125 sweeps per second down to one sweep in five seconds. The sweep-voltage waveshape is the solid sawtooth line shown in Fig. 14-21; the interval from *T1* to *T2* represents the sweep, and from *T2* to *T3* the return of the spot.

In order to permit the convenient analysis of nonperiodic and transient oscillations, provision has been made in the Type 215 for single-sweep operation; the sweep may be initiated either by the operator or by the signal under study.

Description

The equipment comprises a power supply designed to provide the voltages necessary for the proper operation of the unit, a sawtooth voltage generator using a type 6Q5G gas triode, a direct-current amplifier and positioning circuit, a cathode-follower tube (here called an impedance transformer) to couple the sawtooth oscillator to the amplifier, appropriate circuits to allow single-sweep operation, a compensating circuit to insure linearity of the sweep, and a return-trace elimination circuit.

The unit has a number of important advantages over ordinary time-base generators. Nonlinearity of the sweep has been reduced to the point where it is completely negligible at all frequencies. The annoying "electrical backlash" which accompanies many posi-



PARTS LIST

C1	50 μμf 500 v. mica
C2	1 μf 400 v. paper
C3	12-100 μμf trimmer
C4	350 μμf 500 v. mica
C5	350 μμf 500 v. mica
C6	12-100 μμf trimmer
C7	0.1 μf 1000 v. paper
C8	0.1 μf 1000 v. paper
C9	0.001 μf 500 v. mica
C10	8 μf 200 v. paper
C11	1 μf 400 v. paper
C12	0.5 μf 600 v. paper
C13	0.25 μf 400 v. paper
C14	0.05 μf 400 v. paper
C15	16 μf 450 v. elec.
C16	16 μf 450 v. elec.
C17	40 μf 450 v. elec.
C18	40 μf 450 v. elec.
C19	40 μf 450 v. elec.
C20	40 μf 450 v. elec.
C21	0.5 μf 600 v. paper
C22	0.5 μf 600 v. paper
C23	40 μf 450 v. elec.
C24	40 μf 450 v. elec.
C25	40 μf 450 v. elec.
C26	40 μf 450 v. elec.
C27	0.5 μf 600 v. paper
F1	1 amp. 250 v.
L1	8 h. 0.035 amp. 325 Ω D.C.
L2	8 h. 0.035 amp. 325 Ω D.C.
L3	8 h. 0.035 amp. 325 Ω D.C.
L4	8 h. 0.035 amp. 325 Ω D.C.
PL1	Part No. 9-64
PL2	Part No. 9-3
PL3	Part No. 9-1
R1	100 k. ½ w.
R2	2 meg. ½ w.
R3	1.5 meg. ½ w.
R4	1 meg. ½ w.
R5	200 k. ½ w.
R6	150 k. 1 w.
R7	150 k. 1 w.
R8	150 k. 1 w.
R9	150 k. 1 w.
R10	100 k. pot.
R11	100 k. 1 w.
R12	2 meg. 1 w.
R13	2 meg. 1 w.
R14	2 meg. 1 w.
R15	1 meg. 1 w.
R16	2 meg. 1 w.
R17	1 meg. 1 w.
R18	2 meg. 1 w.
R19	2 meg. 1 w.
R20	100 k. pot.
R21	20 meg. ½ w.
R22	100 k. ½ w.
R23	10 k. ½ w.
R24	20 k. ½ w.
R25	12 k. ½ w.
R26	500 k. 1 w.
R27	750 k. 1 w.
R28	4 meg. pot.
R29	400 k. 1 w.
R30	50 Ω 1 w.
R31	250 k. 1 w.
R32	250 k. pot.
R33	20 k. 1 w.
R34	300 k. 1 w.
R35	100 k. pot.
R36	100 k. 1 w.
R37	100 k. 1 w.
R38	500 k. pot.
R39	100 k. 1 w.
R40	100 k. 1 w.
R41	150 k. 1 w.
R42	150 k. 1 w.
R43	150 k. 1 w.
R44	150 k. 1 w.
R45	100 k. 1 w.
R46	100 k. 1 w.
R47	100 k. 1 w.
R48	250 k. 1 w.
R49	250 k. 1 w.
R50	100 k. 1 w.
R51	50 k. pot.
R52	100 k. 1 w.
R53	250 k. 1 w.
R54	250 k. 1 w.
R55	100 k. 1 w.
R56	100 k. 1 w.
R57	100 k. 1 w.
R58	100 k. 1 w.
R59	100 k. 1 w.
R60	100 k. 1 w.
R61	100 k. 1 w.
R62	100 k. 1 w.
S1	S.P. D.T. 60° rotary
S2	D.P. 5T. 30° rotary
S3	S.P. D.T. 60° rotary
S4	S.P. D.T. 60° rotary
S5	D.P. D.T. slide
S6	S.P. S.T. toggle
T1	Part No. 20-32

(K = 1000 ohms)

Courtesy Du Mont Labs.

Fig. 14-20.—Circuit of Du Mont linear time-base generator, Type 215.

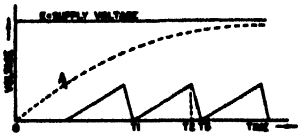


Fig. 14-21.—Sweep-voltage waveshape of the time-base generator, Type 215, is the solid sawtooth line.

Courtesy Du Mont Labs.

tioning circuits has been eliminated. In most single-sweep circuits, the adjustment for single sweep is quite critical, and, when the sweep oscillations have stopped, it is necessary to position the spot off the screen manually. The Type 215, however, is designed so that the adjustment for single sweep and the positioning of the spot off the screen are accomplished simultaneously by simply turning a switch.

The unit has a maximum undistorted balanced output (brought out to terminals at the rear of the cabinet) of approximately 400 volts peak to peak, and when the unit is used in conjunction with the Du Mont Type 175-A Cathode-Ray Oscilloscope, or equivalent, the pattern can, in effect, be spread out in the direction of the sweep to an extent corresponding to approximately three times full screen deflection, or about 15 inches.

Operation of Sweep Voltage Generator

Consider that initially the capacitor *C* in Fig. 14-22 is uncharged. As potential is applied, the capacitor will charge through resistor *R* until the voltage developed across *C* is equal to the firing potential of the 6Q5G gas triode, a tube of the thyratron class, somewhat similar to the type 884 tube. The capacitor will then discharge through the tube until the voltage has dropped to a point where the tube de-ionizes and another cycle, consisting of a gradual rise in voltage across *C* followed by a sudden drop, will begin. The voltage developed across a capacitor charging through a resistance varies with time in the manner indicated by the curve of Fig. 14-21, but if the peak voltage developed across the capacitor is a small fraction of the supply voltage, the corresponding portion of the curve (for example *OA* in Fig. 14-21A) will be essentially linear. The voltage developed across *C* in Fig. 14-22 will closely approximate the sawtooth voltage of Fig. 14-21. Since the charging rate of *C* depends on the product *RC,* the sweep frequency can be varied by varying either *C* or *R*.

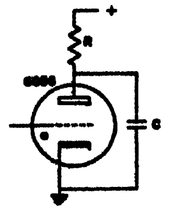


Fig. 14-22.—Plate circuit of sweep voltage generator. The voltage developed across *C* closely approximates the sawtooth voltage of Fig. 14-21.

Courtesy Du Mont Labs.

The operation of the single sweep is as follows: As shown in Fig. 14-23, there is, in parallel with the gas triode and the sweep capacitors, a diode (actually a triode with grid connected to cathode). The cathode of the diode is maintained at a fixed positive potential by means of the voltage divider *R1* and *R2*. When the circuit is to be used to provide a recurrent sweep voltage, the bias of the 6Q5G is kept at such a value that its firing voltage is slightly below the voltage of the cathode of the diode. The highest potential of the diode plate is, accordingly, at no time sufficient to cause the diode to conduct. The diode is, under these conditions, inoperative, and the recurrent sawtooth voltage is developed precisely as described above. For single-sweep operation, on the other hand, the grid of the 6Q5G is given a sufficiently negative bias so that the firing potential of the gas triode is slightly above the potential which must be applied to the plate of the diode to cause it to pass current. Hence, as capacitor *C* charges, a point is reached, before the gas triode fires, at which the resistances *R, R2,* and the plate resistance of the diode form a voltage divider which maintains the potential across the capacitor at a fixed value just slightly below the firing potential of the tube. To produce a single sweep, it is now necessary to apply to the grid of the 6Q5G a positive voltage pulse of sufficient amplitude to cause it to fire; the capacitor then discharges and charges once as described above. Because of the low speed of the sweep and the short discharge time of the capacitors, the time lag between the application of the positive pulse and the beginning of the sweep is not objectionable. The tripping pulse may be applied by the operation of a switch; or the signal under study may be used to initiate the sweep. Under the latter circumstance, it is, of course, necessary that the signal have a positive polarity.

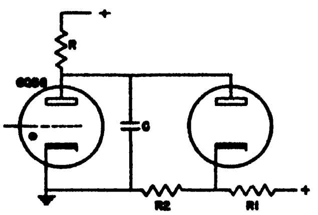


Fig. 14-23.—Single sweep circuit incorporating a diode.

Courtesy Du Mont Labs.

When the sweep is running freely and being used to investigate a periodic signal such as, for example, a sinusoidal wave, the pattern ordinarily will be seen to move across the screen of the cathode-ray tube. This pattern drift results from the fact that the signal frequency is not necessarily an integral multiple of the sweep frequency. The pattern on the screen, accordingly, may start at a different phase after each sweep. However, since the firing potentials of the gas triode depend on the grid voltage, a small portion of the signal voltage fed to the grid insures firing regularly at the same phase of the signal voltage. Since, however, this synchronization voltage adversely affects the linearity of the sweep, it is always advisable that the synchronization voltage be the least necessary for stable lock-in of the pattern.

Blanking Signal

Under certain conditions of operation, as, for example, when it is desired to photograph the pattern on the screen of the cathode-ray tube, the trace caused by the beam during the return time may become objectionable. This return trace is readily eliminated by applying to the modulating electrode of the cathode-ray tube, during the time of the return trace, a negative voltage sufficient to cut off the beam. To provide this blanking voltage, the sawtooth output of the gas triode is amplified and fed to a high-pass filter, whose output is a high-amplitude sharp negative pulse during sweep returns, cutting off the beam, and a very low-amplitude positive voltage during the sweep itself, having no noticeable effect.

Amplifier Circuits

The output from the sawtooth oscillator is directly coupled, through a cathode-follower tube functioning as an impedance transformer, to a combination amplifier and positioning stage which has all the advantages of capacitance coupling and none of its disadvantages. Since direct coupling is used throughout, the annoying backlash in positioning associated with capacitance coupling is avoided. The amplifier circuit is phase-inverting, and the output signal voltage is balanced about ground potential.

Linearity of Time-Base

For a number of reasons, the output of a sawtooth oscillator does not vary linearly with time. This nonlinearity is due partly to the natural curvature (as shown dotted in Fig. 14-21) of the graph of capacitor voltage against time, partly to a small amount of leakage through the sweep capacitors, and partly to inductive impedance of the power supply. The nonlinearity resulting from these causes is particularly noticeable at low sweep frequency. In order, therefore, to insure linearity of sweep in this instrument, there has been incorporated a compensating circuit which feeds back to the plate of the 6Q5G a sawtooth voltage distorted from linearity in a direction opposite to the deviation from linearity of the uncompensated output of the gas triode. The variation from linearity of the combination of these two voltages is completely negligible.

Output Circuit and Cable

The d-c level of the output circuit is brought to zero so that the deflection signal voltages produced will be balanced about ground potential, and hence will not disturb the focus of the oscilloscope with which the generator is used.

The output cable is an integral part of the instrument and should always be used. It allows a convenient direct connection from the Type 215 to the rear terminal strip of the oscilloscope, to which it is connected, without obstructing the front panel controls with unnecessary wires. It also provides the common ground connection. The capacitance of the cable is part of the output compensating circuit, and the instrument has been adjusted at the factory with this cable connected. If it is not used, the linearity of the first part of each sweep-cycle will be adversely affected.

Panel Controls

Any sweep frequency from approximately 0.2 to 125 cps is available in this instrument. The "Frequency Range" switch selects, as the sweep capacitor, one of the capacitors *C10* to *C14* (Fig. 14-20), and so permits the selection of the range of frequencies controlled by the frequency fine control. The approximate ranges controlled by this switch are marked on the front panel, but these are not to be taken as an exact calibration. By varying resistance *R28*, designated "Frequency Vernier," through which the sweep capacitors charge, the sweep frequency may be varied continuously over the range selected by the frequency-range switch. The desired sweep frequency should be obtained as closely as possible with this control in order to minimize the synchronization voltage necessary to stabilize the pattern; for, as pointed out above, the use of too much synchronization voltage will tend to cause nonlinearity of the sweep.

The "Sweep Recurrence" switch selects single or continuous sweeping. When this switch is in the position marked "Continuous," the bias on the grid of the gas triode is such that the single-sweep diode, described earlier, is inoperative, and the sweep oscillations will continue uninterrupted. Throwing this switch to "Single" adjusts the bias on the gas-triode grid so that the diode conducts before the triode fires. A positive voltage pulse is then required on the gas-triode grid to permit a single sweep.

The "Synchronizing Signal Selector," when at the "External" position, makes use of a signal arriving through the corresponding input terminal to trigger the sweep, whether continuous or single. When the switch is turned to the "Manual" position, triggering the sweep is accomplished by momentarily grounding the terminal marked "Manual Trip." This applies through *C9,* a ground voltage pulse to the grid which normally has a negative bias. For remote trip, a jack is provided for connection to the terminal and ground. The two synchronizing methods are actually very similar, except that the manual trip uses an internal positive voltage source, and the external trip requires positive pulses.

The "Time-Base Amplitude Control" determines the amplitude of deflection in the direction of the sweep. When turned up to full gain, and when the unit is used in conjunction with the Du Mont Type 175-A Cathode-Ray Oscilloscope (or equivalent), the pattern on the screen of the cathode-ray tube can be spread out in the direction of the sweep to an extent corresponding to a sweep amplitude which is approximately three times the screen diameter.

Time-Base Position Control

The "Time-Base Position Control" permits adjustments of the spot or pattern in the direction of the sweep. This control makes it convenient to center, for careful study, any portion of the signal under investigation. The position control will permit, when used with the Du Mont Type 175-A Cathode-Ray Oscilloscope, some three inches of positioning each side of center.

When the "Blanking Signal Polarity Switch" is thrown to the minus (−) position, the output from the blanking signal output terminal is a sharp negative pulse, which, when applied to the modulating electrode of a cathode-ray tube, will blank out the return trace. Should it be necessary to amplify the blanking output to obtain a larger blanking voltage, the switch should be thrown to the plus (+) position and the output of the blanking terminal put through a stage of amplification. The resultant output from the amplifier will be of the correct polarity, since there is a phase inversion in amplification.

Factory Adjustments

The following controls are factory adjustments and will rarely require readjustment. They are located on the chassis and will necessitate removal of the chassis from the cabinet for adjustment. These adjustments are:

(1) A d-c level control *R35* which, when properly set, eliminates a change in the zero positions of the sweep, as the sweep amplitude is adjusted.

(2) A linearity adjustment *R10* which adjusts the feedback mentioned earlier, to the proper value for linear sweep. The adjustment is made so that marks, or pips, at equal time intervals, appear equally spaced.

(3) A single-sweep adjustment *R32* which sets the potential at which the single-sweep diode starts to conduct. This control is set as close as reasonably possible to this critical operating point, to preserve sweep linearity.

SWEEP SIGNAL GENERATORS

The sweep signal generator is used in conjunction with the cathode-ray oscilloscope for visual alignment of receiver circuits. It can hardly be considered an accessory to the oscilloscope, but is, rather, a piece of test equipment which requires an oscilloscope for satisfactory interpretation of results.

A sweep signal generator is a device for supplying a signal of rapidly varying frequency. The frequency generally, but not necessarily, increases with time throughout its range, returning abruptly to the starting point and repeating this cycle, or else varying sinusoidally about a definite center frequency. There is also an output, in almost all cases, which provides a voltage proportional to the frequency; this voltage is used as a horizontal sweep supply for an oscilloscope, as will be explained later.

Sweep signal generators are available in audio, TV, AM, and FM ranges, with refinements and special abilities. Audio sweep generators will be discussed in Chapter 15; alignment procedure is also separately treated in Chapter 16. At this time, it is our purpose to show some of the principles involved, and how various commercial generators are designed to achieve the desired results.

Emphasis in this chapter is upon signal generators for television, since these illustrate the various principles more thoroughly. While f-m alignment with the sweep generator is desirable, it is a fact that many f-m sets are so constructed that proper alignment is feasible using a signal generator whose output may be amplitude-modulated only, or even unmodulated.[8]

TV circuits, on the other hand, cannot be aligned properly without a sweep generator, marker system, and oscilloscope.

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[8] Rider, John F., and Uslan, Seymour D., "FM Transmission and Reception," John F. Rider Publisher, Inc., New York, N. Y., 1948.

TV Sweep Generators[4]

The TV sweep generator is a signal generator which furnishes frequency-modulated signals over a number of bands — the TV carrier frequency range, the picture i-f band, and the sound i-f band. In many cases, the same device also furnishes a series of unmodulated r-f signals of predetermined fixed frequency which are within the acceptance band of the TV receiver r-f, picture i-f, and sound i-f systems.

These unmodulated r-f signals are known as marker or calibrating signals. If the TV sweep generator does not furnish these marker signals, then an appropriate external r-f signal is used. Most TV sweep generators offer other facilities, too, which are incidental to the primary purpose of circuit alignment.

The marker signals permit identification of spot points in the swept band and complement the functioning of the sweep generator, Without marker frequencies, proper shaping of the response curve, an operation essential to the proper performance of the tuned circuits in a TV receiver, would be impossible.

Accordingly, the function of the frequency-modulated signal source is to furnish a test signal over a band of frequencies whereby it is possible to adjust the appropriate receiver circuits for correct operation — that is, correct bandpass characteristics.

In practice, the TV sweep generator is used in conjunction with an oscilloscope, to develop a visible response curve. The marker generator furnishes the signals which "ride" on the response curve and permit identification of the band limits and points between them. Both the sweep generator and marker generator can be used with devices other than an oscilloscope, but because of a continuously visible indication, the oscilloscope is by far the most convenient indicator.

Frequency Requirements

In Fig. 14-24 are shown three radio transmission channels, one for standard a-m broadcasting, one for f-m broadcasting, and one for television. The over-all frequency spread allowed the stations is indicated. For amplitude modulation, it is about 10 kc, but may be as much as 20 kc, depending on the transmitter and location. For frequency modulation, the channel bandwidth is 200 kc, although only ±75-kc deviation represents 100 per cent modulation; a 25-kc guard band is allowed on each side of the ±75-kc peak deviation from the center frequency. Television transmission requires much wider channels, each station being allowed 6 Mc.

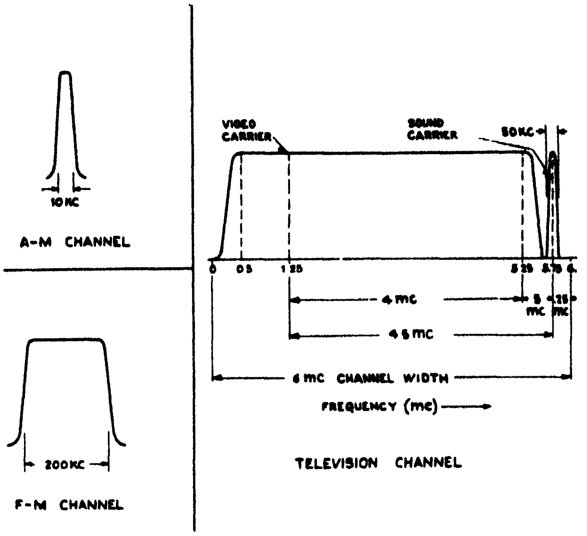

Fig. 14-24.—Three radio transmission channels, one for standard a-m broadcasting, one for f-m broadcasting, and one for television. The over-all frequency spread allowed the stations is indicated.

Checking Wideband Circuits (TV)

The most convenient method of checking the performance of a wideband system, such as a TV amplifier, is to use a sweep generator and a cathode-ray oscilloscope, for development of a response curve. However, because of the general structure of the TV broadcasting system and the design of the receivers, it is not enough to view the response curve for one particular channel.

The response of the circuits to frequencies of adjacent channels must be checked, if for no other reason than to make certain that only the desired signals will be received and that all others will be rejected; otherwise, interference will be experienced. Therefore, any signal source intended to provide a frequency-modulated signal for such circuit testing must furnish signals over a frequency band far wider than the individual station-channel under test. We are now concerned with TV sweep generators; the statement applies, however, with equal force to a-m and f-m channels.

Sweep-Generator Fundamentals

Obviously, all sweep generators must possess a fundamental oscillator. Since the final output is an f-m signal of constant amplitude, additional circuits are required to produce the frequency changes.

To attain a change in frequency of from 5 to 10 Mc above and below a center frequency of from 100 to 200 Mc is not difficult. It often happens that undesired variations in frequency, of this magnitude, are caused by shifts in the operating potentials. But when it is necessary to secure a frequency swing of from 10 to 15 Mc around a center frequency about 25 Mc, such as may be required for checking picture i-f channels, or a swing of perhaps 2 Mc when the center frequency is 4.5 Mc, needed for checking sound i-f channels in intercarrier systems, convenient operation by direct sweeping or frequency changing in the oscillating circuit is virtually impossible. Recourse to the beat-frequency or heterodyne method is then required.

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[4] Rider, John F., "TV sweep generators," *Radio Service Dealer*, vol. 9, p. 11, November, 1948. The following material is an abridged excerpt from this article.

Heterodyne Technique

If a larger frequency deviation is required, a high center frequency must be used to keep the percentage deviation low for practical reasons. Heterodyne methods can then be used to reduce all frequencies by a fixed amount, so that the same number of Mc deviations may be made to appear about any desired center frequency. This is, obviously, the familiar frequency conversion found in superheterodyne receivers.

To perform the conversion, two signals of different frequency are applied to a nonlinear impedance. The resulting output consists of the fundamentals, sum and difference frequencies, and harmonics. By means of a selective output circuit, the desired signal is chosen. The nonlinear impedance may be a vacuum tube or crystal rectifier.

If one of the signals is frequency-modulated, and it is heterodyned by an unmodulated signal, an f-m signal output will be obtained with a center frequency equal to the sum or difference of the original f-m center frequency and heterodyne signal frequency. In this system, accordingly, wideband sweep signals can be developed.

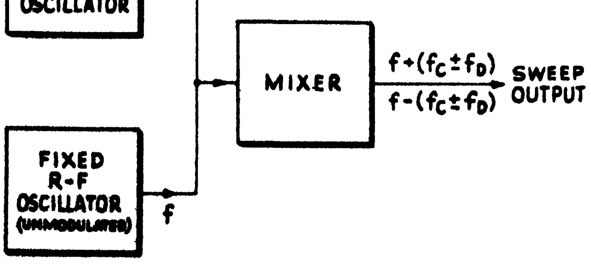


Fig. 14-25.—Block diagram of heterodyne technique. By means of the heterodyning process, either a reduction or an increase in the center frequency can be accomplished without disturbing the amount of deviation.

An example to illustrate the above point is given in Fig. 14-25. The f-m oscillator produces a signal with a center frequency f_C. The frequency deviation is $\pm f_D$. For the sake of discussion, assume f_C to be 100 Mc and f_D to be ±5 Mc. Therefore, the frequency limits are $100 - 5 = 95$ Mc, and $100 + 5 = 105$ Mc. The deviation is $105 - 95 = 10$ Mc.

Let the frequency of the fixed r-f oscillator be 80 Mc. As can be seen from Fig. 14-25, the two signal sources feed into a mixer. The mixer output will contain, as the limits of the frequency variation:

$$95 + 80 = 175 \text{ Mc}$$
$$95 - 80 = 15 \text{ Mc}$$

and

$$105 + 80 = 185 \text{ Mc}$$
$$105 - 80 = 25 \text{ Mc.}$$

It will also contain other signals, but our principal interest is in the sum and difference frequencies produced. We have, then,

$$185 - 175 = 10 \text{ Mc and}$$
$$25 - 15 = 10 \text{ Mc.}$$

Using the heterodyne system and an original deviation of 10 Mc we have obtained an output which likewise has a deviation of 10 Mc. However, the center frequencies now are

$$\frac{(185 - 175)}{2} + 175 = 180 \text{ Mc}$$

and

$$\frac{(25 - 15)}{2} + 15 = 20 \text{ Mc.}$$

By means of the heterodyning process, either a reduction or an increase in the center freqeuncy of an f-m signal can be accomplished without disturbing the amount of deviation.

The figures quoted are purely illustrative. Other unmodulated and f-m signal frequencies could be and are used in sweep generators. The use of the heterodyne system is common to all TV sweep generators as a means of developing all or certain output signals, although it is necessary to mention that this "beating down" arrangement is used in most instances for the generation of the wide-swing low center-frequency signal required for i-f and associated circuit alignments.

The heterodyning system is not generally used for the development of the f-m signal at the TV carrier frequencies. As the ratio of the center frequency to the deviation increases, the design of direct sweeping circuits becomes simpler, because the *percentage* deviation is less. This is true for the high frequencies of the TV carrier.

Obtaining the Sweep

To obtain the deviations required in either the direct or heterodyne systems, either mechanical or electrical methods may be used.

Mechanical Methods

Let us examine, to start with, an ordinary *L-C* resonant circuit, shown in Fig. 14-26. The earliest form of frequency modulation probably occurred with the swinging of an antenna coupled to an r-f oscillator-type transmitter. In effect, *C* in Fig. 14-26 was paralleled with a variable capacitor, which changed the frequency.

One of the earliest forms of sweep-frequency generators used a small variable capacitor, in parallel with the main oscillating circuit capacitor, which was continuously rotated by a motor. In present-day mechanical systems, a vibrating mechanism somewhat similar to a dynamic loudspeaker assembly is made to vary either the capacitance or the inductance of the f-m oscillator.

If capacitance variation is employed, the vibrating mechanism, actuated by a variable voltage 60-cps signal, may be mounted on a specially constructed variable capacitor which governs the tuning of an f-m oscillator system. The capacitance increases or decreases around a center reference value in accordance with the variations in amplitude of the driving voltage.

The driving voltage may be called the sweep amplitude control, since a high value of driving voltage causes a great displacement of the vibrating mechanism and, therefore, a great change in frequency, or wide sweep. Small amounts of driving voltage, on the other hand, produce less mechanical change and reduced frequency variation or narrower sweep. Inductance variations may be obtained in a somewhat similar manner to cause frequency changes in a tuned circuit. A metal plate located near the f-m oscillator coil is vibrated, affecting the inductance, and a frequency modulation then occurs, since variation of either *L* or *C* in any tuned circuit will change the resonant frequency.

This change, as in the previous case, is, within limits, proportional to the amplitude of the driving voltage acting on the electromechanical system.

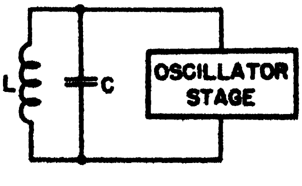


Fig. 14-26.—Basic *L-C* resonant circuit.

Electronic Methods

There are also means of varying frequency directly by varying electrical quantities. The reactance-tube modulator and the variable-frequency klystron are two common means of doing this.

A. In the reactance-tube modulator, an electron tube is used as a variable inductance or capacitance in shunt with the tuned circuit of an oscillator. Voltage changes at the grid of the tube, in accordance with the modulating signal variations, produce the required frequency modulation. The possible amount of direct-frequency modulation is somewhat limited, and this system is generally used with a beat oscillator to obtain a wideband sweep. The theory of f-m oscillators is given adequately in standard texts and, therefore, will not be repeated here.

B. In klystron sweep generation, two reflex klystrons are used. The klystron is a special type of tube which includes the oscillator circuit. This circuit can be mechanically adjusted over a small range to set the frequency, and a fine adjustment can be made by varying the electrode voltages properly. One klystrons functions as a signal oscillator in the neighborhood of 10,000 Mc; it can be varied in frequency.

The other klystron is the f-m oscillator. This tube also operates at a frequency of approximately 10,000 Mc, but by varying the voltage applied to one of the tube elements, the repeller plate, the frequency of the output signal is thereby varied correspondingly, within certain limits.

These frequencies, however, are entirely too high for normal use, and the heterodyning principle is applied to produce output signals having frequencies in the usable region. The reason for using these high frequencies is, as explained before, to have a low *percentage* deviation with an adequate *absolute* (number of cycles) deviation.

Marker Generation

The sweep generator would be limited in usefulness without the marker generator. This is a means of giving increased or decreased amplitudes for certain known frequencies, so that they may be recognized on the face of the cathode-ray oscilloscope. Two methods of marker identification are in common use. One of these is the generation of unmodulated r-f signals in one or more crystal oscillator circuits. The signal frequency is known from the calibration of the device.

When the signal is injected into the circuit being checked, it appears as a "marker pip," or a few cycles riding on the response curve of the system being viewed on the oscilloscope screen. Exaggerated examples are shown in Fig. 14-27. In some units, this marker signal source is a part of the TV sweep-generator assembly. In others, it is a separate unit.

The second system uses an absorption-type wavemeter which creates a dip in the response curve at a point corresponding to the resonant frequency of the wavemeter. The dip width is narrow because of the high Q of the wavemeter and loose coupling to the circuit under test.

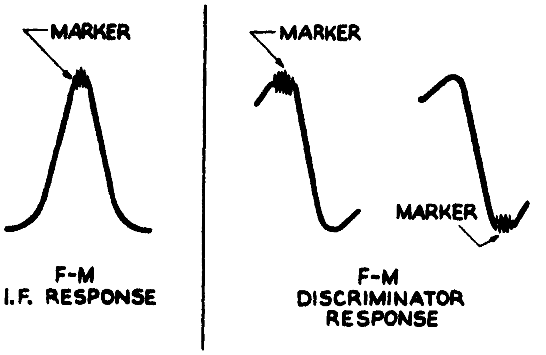


Fig. 14-27.—Exaggerated examples of marker-generator signal indications.

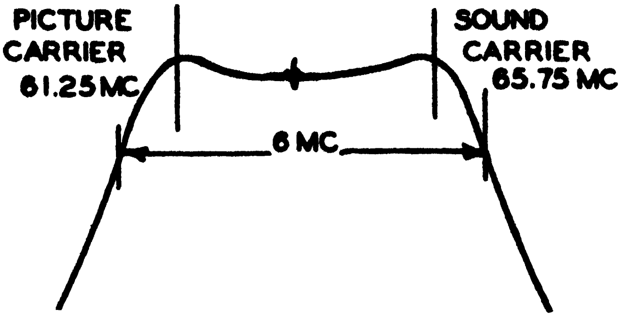
The range of marker frequencies provided in such calibrating systems extends over the TV carrier frequency band and over the sound and picture i-f ranges, including the sound-trap frequencies.

The marker signals must be available over the TV carrier range and cover the alignment frequencies of the i-f channels in the TV receiver.

While the principal purpose of the marker generator is to identify points on the response curve, as indicated on a cathode-ray-tube screen, it is also of value as a nonsweeping signal generator. Crystal-controlled oscillators will furnish a great many fixed points, accurately calibrated, usually in multiples of a quarter-megacycle; variable-frequency oscillators can be adjusted to any value by means of a properly graduated dial. Instruments are available in which either crystal-controlled or manually adjusted operation may be obtained.

Circuit Alignment and Response Curves

Although the process of alignment and the shape of the response curves will be discussed at length in Chapter 16, some mention of these subjects must be made at this time, so that the application of the sweep generator may be understood.



Courtesy RCA

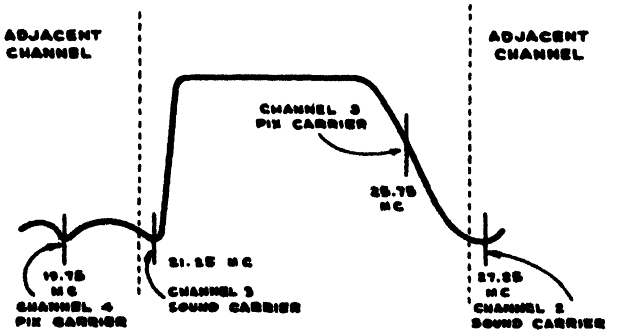
Fig. 14-28.—Response curve for the r-f stages of a typical television receiver.

Alignment is a process of adjusting the resonant circuits of a system so that the individual and the over-all frequency responses will be as desired. The process is one of injecting a signal into a stage of amplification and noting the corresponding output; if this is done at many frequencies for each stage, its frequency characteristic may be plotted, giving the response curve.

The sweep-frequency generator applies frequencies over the entire band under study; the output, when applied to an oscilloscope after detection, will produce the response curve if the oscilloscope sweep follows the frequency sweep.

Fig. 14-28 shows a response curve for the r-f stages of a typical television receiver which, for the purpose of illustration, is tuned to channel 3. There is a 4.5-Mc separation between the sound and video signal carriers. Since the r-f portion of the receiver must pass both the sound and video signals, a bandpass of approximately 6 Mc is required.

The ideal response curve for the picture i-f channel is shown in Fig. 14-29. In this case, the receiver uses a picture i-f of 25.75 Mc for channel 3. The frequency relation of sound carrier to picture carrier is reversed in the i-f amplifier, since the local oscillator operates at



Courtesy RCA

Fig. 14-29.—Ideal response curve for the picture i-f channel.

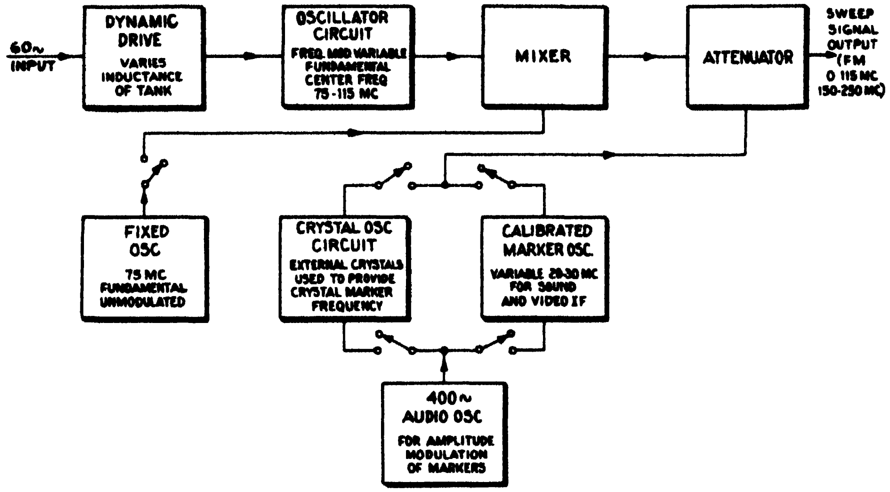

60~ INPUT
DYNAMIC DRIVE
VARIES INDUCTANCE OF TANK
OSCILLATOR CIRCUIT
FREQ. MOD VARIABLE FUNDAMENTAL CENTER FREQ 75-115 MC
MIXER
ATTENUATOR
SWEEP SIGNAL OUTPUT (FM 0 115 MC 150-250 MC)
FIXED OSC
75 MC FUNDAMENTAL UNMODULATED
CRYSTAL OSC CIRCUIT
EXTERNAL CRYSTALS USED TO PROVIDE CRYSTAL MARKER FREQUENCY
CALIBRATED MARKER OSC.
VARIABLE 20-30 MC FOR SOUND AND VIDEO IF
400~ AUDIO OSC
FOR AMPLITUDE MODULATION OF MARKERS

Fig. 14-30.—Block diagram of a commercial television sweep generator which uses mechanical variation of an inductance.

a higher frequency than the carrier. The picture channel will ordinarily have a bandpass of from 3 to 4 Mc wide, the wider band giving increased picture definition.

The picture i-f response may vary with change in gain, and for this reason, the circuits should be aligned with approximately the same signal input and gain-control setting as they will receive under normal operating conditions. Because many of the signal generators have attenuators in the various output circuits, it is possible to align the television receiver at any point from minimum to maximum contrast. The relative gain per stage can be accurately estimated on the screen of the oscilloscope.

Alignment of any portion of a television receiver should not be attempted unless there is reasonable evidence pointing toward misalignment. The sweep generator can be employed to check the condition of alignment, as well as to facilitate the process of aligning. As mentioned earlier in this chapter, the marker generator, used extensively in conjunction with sweep generators, produces on the oscilloscope trace one or more small marks or pips of known frequency, so that not only the shape of the response curve, but also its position in the frequency spectrum, can be determined.

PRACTICAL SWEEP-GENERATOR CIRCUITS

The principles used in sweep generators for test and adjustments of TV and other circuits are combined in many ways. A few representative examples will be discussed at some length, according to the commercial importance or technical interest. When there is a special marker oscillator, designed specifically for a sweep generator, its description is included herewith.

Variable-Inductance Multirange Sweep Generator

Fig. 14-30 is a block diagram of a commercial TV sweep generator which uses mechanical variation of an inductance. The arrows on the connecting lines indicate the direction of passage of the signals in the system. The 60-cps input is the sweep-control voltage which actuates the vibrating plate for inductance variation in the f-m oscillator circuit.

A point of special interest is the method of selecting frequency ranges. The f-m difference-frequency signals in the output are obtained by using different combinations of fundamental and second-harmonic frequencies of the two oscillators.

This unit has four f-m ranges — 0-40 Mc, 35-75 Mc, 75-115 Mc, and 150-230 Mc. The 0-40 Mc range is obtained by means of the difference between the fundamental frequencies of the two oscillators, the 35-75 Mc range is derived by means of the second harmonic of the fixed oscillator beating against the fundamental of the f-m oscillator, the 75-115 Mc band is obtained from the f-m oscillator only, and the 150-230 Mc band is obtained by using the second harmonics of the f-m oscillator alone.

Two marker oscillators are used. One uses crystals which may be selected for different fixed frequencies, while the other is a variable-frequency type in the 20-30 Mc range and is used for i-f alignment.

RCA WR-59A Television Sweep Generator

Another example of a TV sweep generator, the RCA WR-59A, is shown in block diagram form in Fig. 14-31; the full circuit diagram is given in Fig. 14-32. F.m. is obtained by means of a vibrating capacitor in the oscillator circuit. The capacitor is driven by

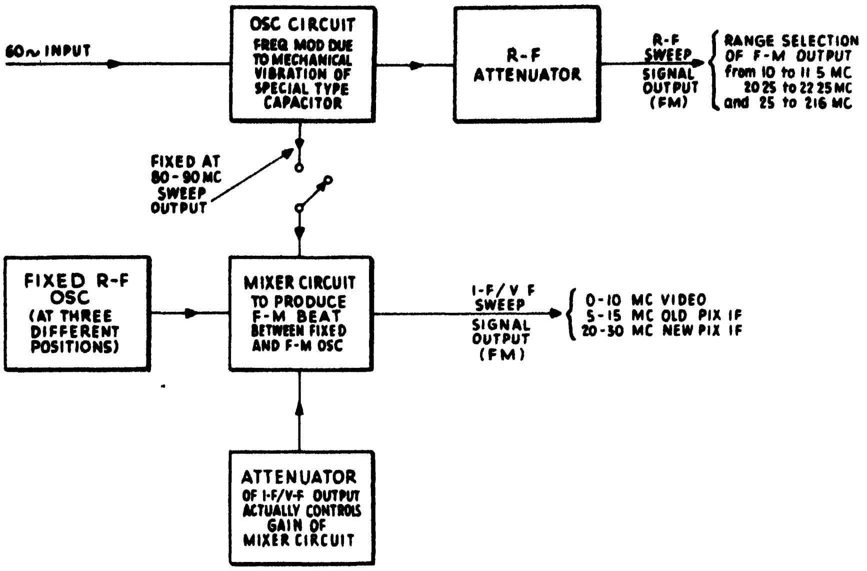

60~ INPUT
OSC CIRCUIT
FREQ MOD DUE TO MECHANICAL VIBRATION OF SPECIAL TYPE CAPACITOR
R-F ATTENUATOR
R-F SWEEP SIGNAL OUTPUT (FM)
RANGE SELECTION OF F-M OUTPUT from 10 to 11 5 MC 20 25 to 22 25 MC and 25 to 216 MC
FIXED AT 80-90 MC SWEEP OUTPUT
FIXED R-F OSC (AT THREE DIFFERENT POSITIONS)
MIXER CIRCUIT TO PRODUCE F-M BEAT BETWEEN FIXED AND F-M OSC
I-F/V F SWEEP SIGNAL OUTPUT (FM)
0-10 MC VIDEO
5-15 MC OLD PIX IF
20-30 MC NEW PIX IF
ATTENUATOR OF I-F/V-F OUTPUT ACTUALLY CONTROLS GAIN OF MIXER CIRCUIT

Fig. 14-31.—Block diagram of RCA WR-59A television sweep generator.

a 60-cps voltage. The f-m signal is fed to a mixer, with a signal from an adjustable-frequency oscillator.

The f-m oscillator output signal is available through its own attenuator in the sound i-f channel range and TV carrier range. The output for the video or picture channels is taken from the mixer tube, and attenuation control of the output signal level is obtained in the mixer stage. This particular generator requires a separate marker-frequency unit, the RCA WR-39A Television Calibrator. Maximum convenience of operation is provided by the circuit design; tuning is not required — switching permits selection of all the desired signal frequencies.

Radio-Frequency Oscillator

The r-f portion of the circuit utilizes a 6J6 (*V1*) as a push-pull oscillator which is directly frequency-modulated at power-line frequency by the mechanical motion of the capacitor plates *C10* and *C11*. These plates are driven by a precision moving-coil type drive assembly. The moving element of the capacitor is suspended by corrugated beryllium-copper straps. The degree of f.m. is determined by the "Sweep Width" control, which sets the voltage applied to coil *L22* of the driver. The channel frequency of the r-f oscillator is determined by the setting of the "Channel" switch. The signal from the r-f oscillator is taken from the "R-F Output" jack on the front panel of the instrument. A piston-type variable-capacitor, "R-F Attenuator," in the r-f output circuit controls the level of the output signal.

In order to produce the zero baseline, 6AT6 square-wave generator *V5* biases the r-f oscillator to cutoff for one-half of the modulation cycle. The r-f output from the sweep generator will, therefore, sweep in one direction, and the output will be zero for the return trace. The blanking-phase control *R60* adjusts the phase of the square wave so that the r-f oscillator will commence its operation at minimum frequency and cease operating at the maximum-frequency point.

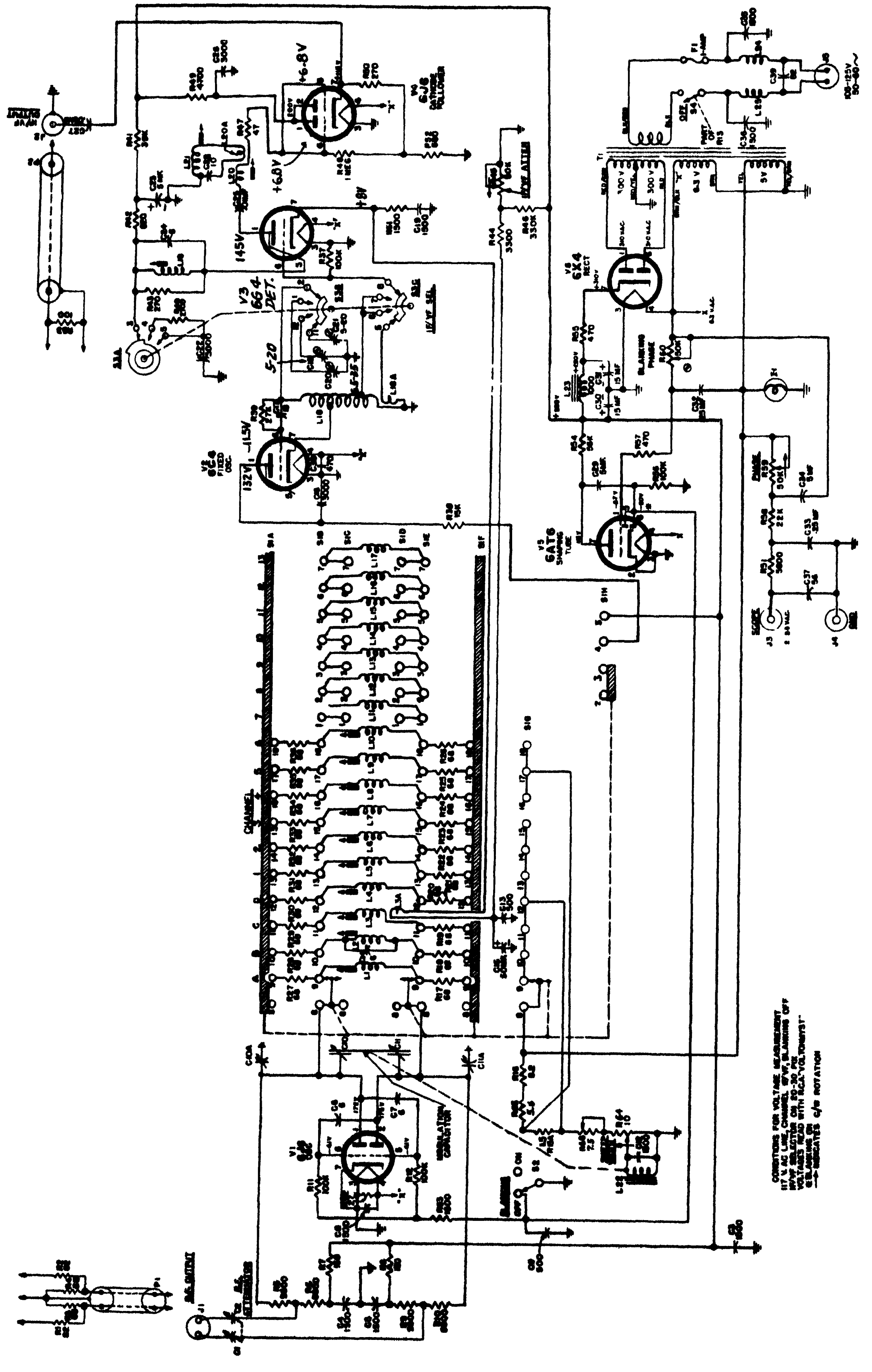
The a-c voltage at the "Scope" and "Ground" terminals produces a sine-wave sweep when applied to the horizontal-deflection circuit of a cathode-ray tube. The "Phase" control phases the horizontal sweep with respect to the output from the Sweep Generator.

IF/VF Ranges

When the "Channel" selector is set to position *C*, the output from the r-f oscillator is fed to a 6C4 heterodyne-detector tube *V3*. The output from a 6C4 fixed oscillator *V2* is also fed to the detector circuit, and the difference frequency produced by the heterodyne action is utilized for the IF/VF signals.

The variable r-f oscillator, when the "Channel" selector is in position *C*, sweeps from 80 to 90 Mc. The frequency of the fixed oscillator is determined by the setting of the "IF/VF Sel" switch. Putting this switch in the "Video 0-10" position tunes the fixed oscillator to 80 Mc, which, when heterodyned with the r-f signal of 80-90 Mc, will result in 0-10 Mc signal output. Also, at this setting of the switch, an absorption circuit is made effective to produce a marker dip in the oscilloscope trace at 3 Mc. The "IF/VF Sel" switch when tuned to "PIX 5-15" or "PIX 20-30" will give an output of 5-15 or 20-30 Mc, respectively.

The output from the 6C4 detector circuit is fed to a



Courtesy RCA

Fig. 14-32.—Circuit of RCA WR-59A television sweep generator.

6J6 cathode follower *V4*. The IF/VF output is taken from the cathode of the 6J6. The "IF/VF Atten," which controls the IF/VF signal level, acts as a gain control in the cathode circuit of the 6C4 detector.

RCA WR-39 Television Calibrator

The WR-39A Television Calibrator is a marker-frequency generator, designed to be used in conjunction with the WR-59A Sweep Generator. The instrument combines the precision of crystal oscillators with the versatility of the variable-frequency oscillator. Incorporated in the instrument are all the elements of a complete frequency-measuring system, a variable-frequency oscillator, two crystal oscillators, a heterodyne detector, and an audio amplifier to aid in determining zero beat.

To permit identification of the frequencies of very weak signals, a phone jack is provided on the front panel. A jack is also provided for the introduction of an external modulating voltage of any frequency when modulation of the output is desired.

General Electric Sweep Generator Type ST-4A

The General Electric Sweep Generator Type ST-4A is designed to produce an exceptionally wide linear sweep for use in television and f-m alignment and testing. The circuit is shown in Fig. 14-33.

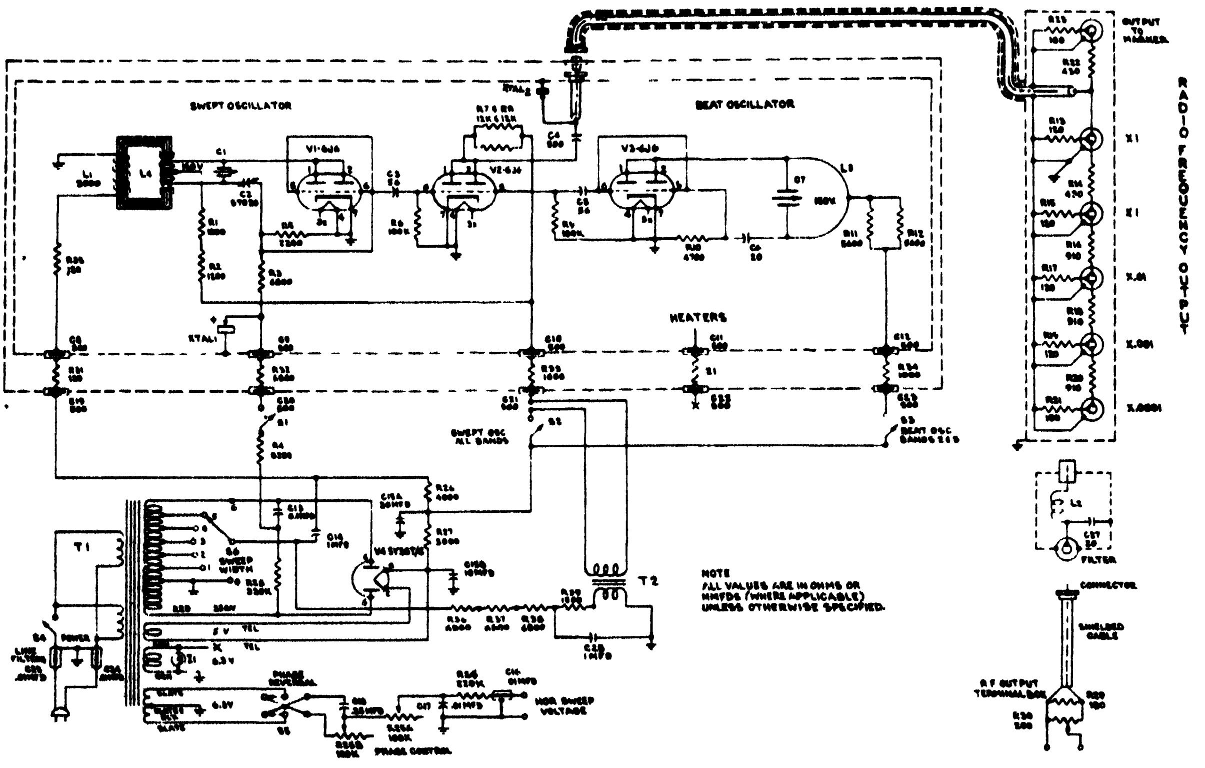
Circuit and Frequencies

There are two oscillators, a linear sweep tunable through a range of 170 to 220 Mc, with adjustable sweep width, and a nonsweeping oscillator tunable from 220 to 275 Mc; there is also a mixer stage. Either of the two oscillators may be operated alone to obtain an output in the stated frequency range, or both may be used together to obtain a beat-frequency output for i-f alignment.

Linear Sweep

The linear sweep is produced by electrical means, directly, and therefore, has no moving parts. A 6J6 oscillator has its coil wound around a powdered iron core placed in the magnetic circuit, or field, of a laminated iron reactor. The reactor has 60-cps voltage applied to its winding; the resulting flux varies the permeability of the powdered iron core, and therefore, the inductance of the coil. Thus the oscillator output is swept back and forth at power line frequency. Sweeps of from 500 kc to more than 15 Mc are available by applying more voltage from the transformer secondary to the reactor winding with the "Sweep Width" switch.

The high-frequency band (170 to 220 Mc) is covered by the Swept Oscillator alone, providing a maxi-



Courtesy General Electric

Fig. 14-33.—Circuit of the General Electric sweep generator Type ST-4A.

Courtesy General Electric

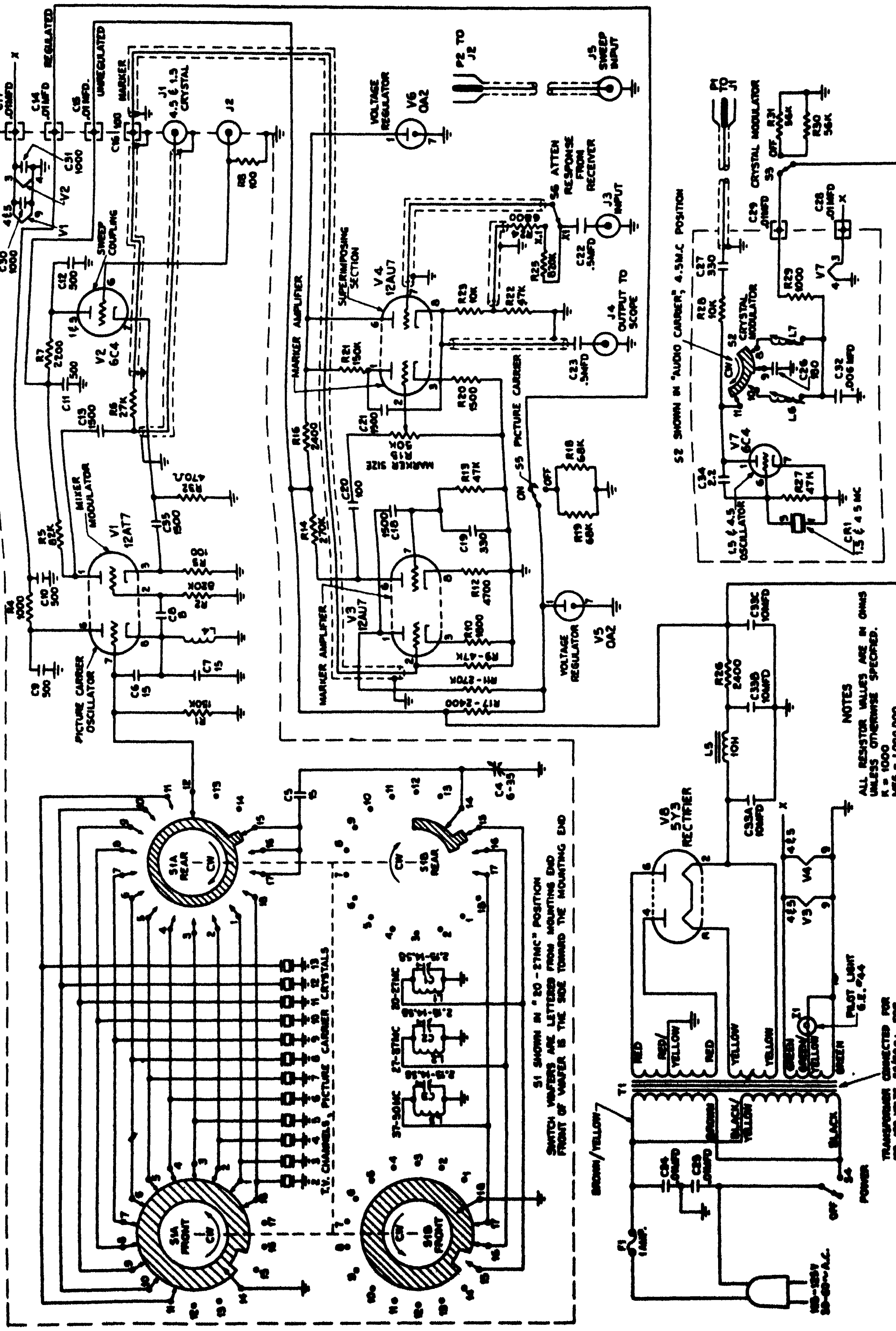

V1 12AT7
MIXER MODULATOR
PICTURE CARRIER OSCILLATOR
V2 6C4
SWEEP COUPLING
V3 12AU7
MARKER AMPLIFIER
V4 12AU7
SUPERIMPOSING SECTION
MARKER SIZE
V5 OA2
VOLTAGE REGULATOR
V6 OA2
VOLTAGE REGULATOR
V7 6C4
4.5 MC CRYSTAL OSCILLATOR
CRYSTAL MODULATOR
V8 5Y3 RECTIFIER
J1 4.5 & 1.5 CRYSTAL
J2
J3 INPUT
J4 OUTPUT TO SCOPE
J5 SWEEP INPUT
P2 TO J2
P1 TO J1
S5 PICTURE CARRIER
S6 ATTEN RESPONSE FROM RECEIVER
S2 SHOWN IN "AUDIO CARRIER", 4.5 M.C POSITION
REGULATED
UNREGULATED
MARKER
S1A REAR
S1B REAR
S1A FRONT
S1B FRONT
T.V. CHANNELS 2-13 PICTURE CARRIER CRYSTALS
S1 SHOWN IN "20-27MC" POSITION
SWITCH WAFERS ARE LETTERED FROM MOUNTING END FRONT OF WAFER IS THE SIDE TOWARD THE MOUNTING END
NOTES
ALL RESISTOR VALUES ARE IN OHMS UNLESS OTHERWISE SPECIFIED.
K = 1000
MEG. = 1,000,000
ALL CAPACITOR VALUES ARE IN MICRO-MICROFARADS UNLESS OTHERWISE SPECIFIED.
MFD = MICROFARADS
TRANSFORMER CONNECTED FOR 105-125 VOLTS 50/60~. FOR 210-250 VOLT OPERATION, CONNECT BROWN TO BLACK/YELLOW & A.C. LINE TO BLACK & BROWN/YELLOW
PILOT LIGHT G.E. #44
F1 1 AMP.
105-125V 50-60~ A.C.
S4 POWER
OFF
T1

Fig. 14-34.—Circuit of General Electric marker generator Type ST-5A.

mum output of at least 0.5 volts. The oscillator output is fed to a mixer tube and then to the r-f attenuator network, and the equipment output connections.

The intermediate frequencies are covered by beating one oscillator against the other, which is tuned to a fixed frequency. Both oscillator outputs are fed to the mixer tube and the difference frequency signals of 4 to 110 Mc appear at the output jacks. In this range the maximum voltage is at least 0.1 volt. A removable filter is included for use at the lower frequencies to attenuate unwanted high frequencies, particularly the oscillator signals when the beat frequency is desired.

The output attenuator has five steps with a vernier control located at the terminal end of the output cable. This arrangement allows complete control from maximum output to below 20 microvolts.

Phasing Control and Other Features

Phasing control is provided so that the sweep will not start at an intermediate point, of the band being swept, on the horizontal sweep output for the oscilloscope. The sweep voltage is taken from separate windings on the power transformer and fed through a circuit which allows a total phase shift of more than 360° for complete flexibility.

Other particular circuit features of this instrument include a cutoff pulse network, which stops the sweep oscillator during the return trace on the oscilloscope screen; in this way, the return sweep can be used as a zero reference line. There is also a slight unwanted amplitude modulation of the output of the sweep-frequency oscillator, which would cause a distortion of the trace on the oscilloscope; a special network *R36, R37, R38, R39,* and *C28* applies a voltage of proper phase and amplitude through the transformer *T2* to the oscillator, for the remedy of this shortcoming.

The equipment is completely enclosed in double metallic shielding with double filters on all leads entering the oscillator compartment. Thus, signal leakage is kept to a value that allows observation of response of the most sensitive television receiver, set at maximum gain, without overloading.

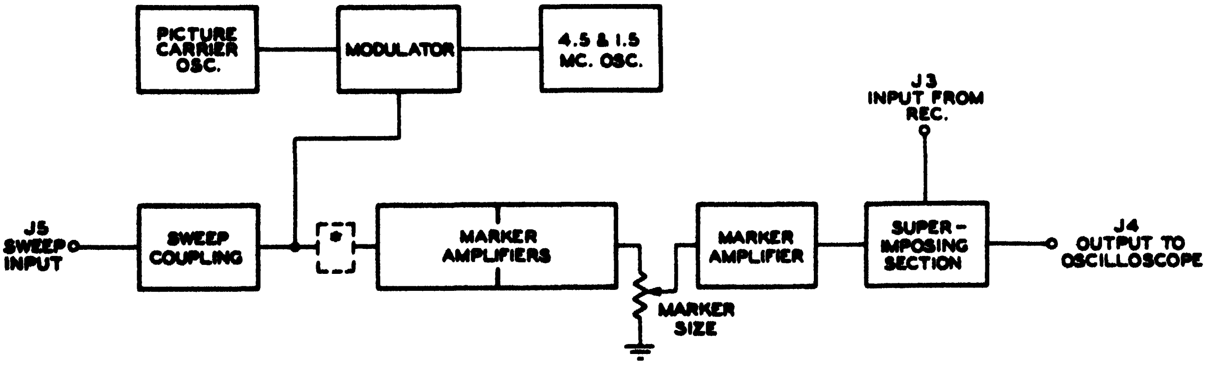
Marker Generator ST-5A

The companion unit for the GE Type ST-4A is the Marker Generator Type ST-5A, whose circuit diagram is shown in Fig. 14-34. The marker generator uses two oscillators, a modulator, a three-stage amplifier, and two mixing stages to produce markers necessary to the alignment of f-m and television receivers.

The block diagram in Fig. 14-35 shows the general circuit layout of the equipment. It will be noticed that there is no r-f output from the generator. The markers are not introduced into the receiver, but superimposed on the receiver output. Therefore, there is no distortion of the response curve and trap circuits can be tuned without losing the markers.

Picture Carrier Oscillator

The picture carrier oscillator uses half of a 12AT7 tube *V1*. It is a modified Colpitts type with a 15-position gang switch to switch any one of twelve crystals or three tunable circuits into its grid circuit. For greater frequency stability, the crystal fundamental frequencies are kept low and harmonic outputs are used for the desired picture carrier frequencies. The output of the picture carrier oscillator is taken from the cathode and fed to the grid of the mixer modulator tube, which is the other half of *V1*.



* ELECTRICALLY THE MODULATOR TUBE ACTS AS AN UNTUNED DETECTOR IN THE CIRCUIT.

Courtesy General Electric

Fig. 14-35.—Block diagram of General Electric marker generator Type ST-5A.

Sweep Coupling

A sampling of the sweep-generator output is cabled to the marker generator. This voltage is applied to the grid of the sweep coupling tube, which prevents interaction between the pieces of equipment. The output of the tube is fed from its cathode to the cathode of the mixer modulator stage through *C35*. Since the mixer modulator grid is being driven by the picture carrier oscillator across a large value grid resistor, it is biased to the nonlinear portion of its operating curve. Thus, the tube acts as an untuned detector and demodulates the combined signals applied to it. This output appears at the plate-load resistor and is taken off through *C13* to the grid of the first marker amplifier stage *V3*.

Marker Amplifiers

The three-stage marker amplifier, including *V3* and half of *V4*, has been designed to give almost no response to a 60-cps square wave. This is necessary because, when blanking is used, the sweep operates only during the trace period and is turned off for the retrace period. Therefore, unless the design is such as to prevent it, the detected output level would shift at the sweep rate of 60 cps.

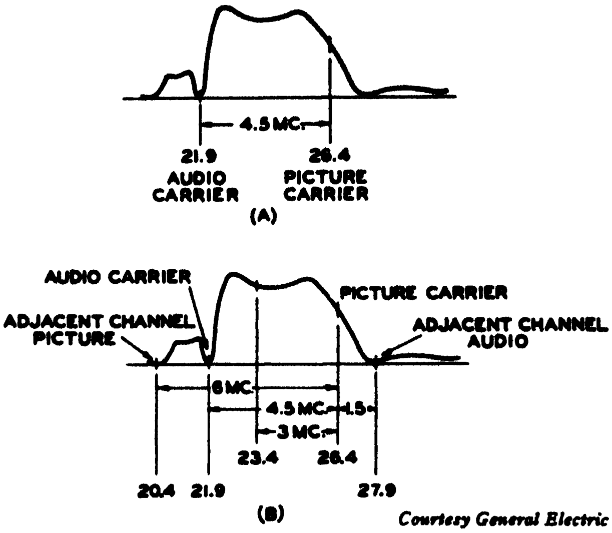
The control of the marker-amplifier output is accomplished by the marker size control, in the grid circuit of the last marker amplifier. The output of the final marker-amplifier tube is fed to the cathode of the superimposing section.

Superimposing Section

The output of the receiver being aligned is connected to *J3*, which injects the signal on the grid of the second half of *V4* in the superimposing section. The detected marker signals are fed to the cathode of the same tube and the combined output appears at *J4* to be transferred to the vertical input circuit of the test oscilloscope. The receiver response may be reduced to one-tenth of its value by turning the attenuator switch *S6*.

4.5 and 1.5 Mc Oscillator

The crystal used in the crystal oscillator grid circuit is triple cut, and normally oscillates at 4.5 Mc. By changing the plate tuning, it will also oscillate at 1.5 Mc. The plate-load switching is done with the "Crystal Modulator Selector Switch" *S2*. B+ can be removed by actuating the "Crystal Modulator On-Off Switch" *S3*.



Courtesy General Electric

Fig. 14-36.—Adjacent channel and band-pass response are illustrated by these curves.

The output of the crystal modulator is fed to the plate of the modulator tube *V1*, where it beats against the picture carrier oscillator output, producing markers ±4.5 Mc to show the relation of picture carrier to audio carrier (Fig. 14-36).

In the 1.5-Mc position, markers are produced every 1.5 Mc across the response curve, allowing the operator to check adjacent channel response and band-pass characteristics (Fig. 14-36).

The Kay "Megasweep"[5]

Because it is not of a usual type, since it uses klystrons instead of the more common tubes, the Kay "Megasweep" sweep-frequency signal generator deserves special discussion. The general principles of klystron oscillators and signal generators were given earlier in this chapter.

Oscillator Components

Fig. 14-37 gives the block diagram of the sweeping oscillator, which is a heterodyne-type generator combining the outputs of two X-band (3-cm) reflex klystrons of the Sheppard-Pierce type in a mixing network. One klystron, called the signal oscillator, is at a fixed frequency during operation of the instrument. The other, designated the local oscillator, is swept in frequency by the application of a sawtooth wave to its repeller. The two outputs are heterodyned in a microwave crystal mixer and the resulting difference frequency made available at the output terminals.

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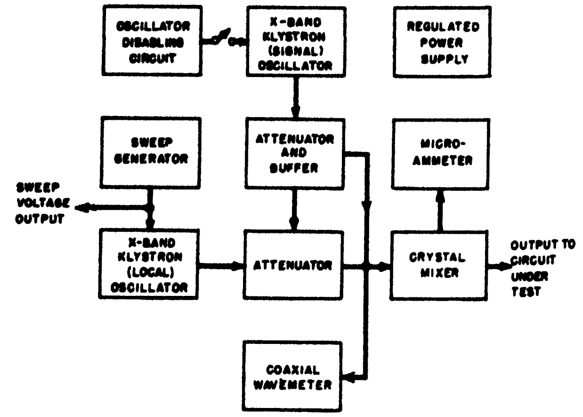
[5] Engineering staff, Kay Electric Co., Newark, N. J., "Wide range sweeping oscillator," *Electronics*, vol. 20, p. 112, August, 1947. The following material is an abridged excerpt from this paper.



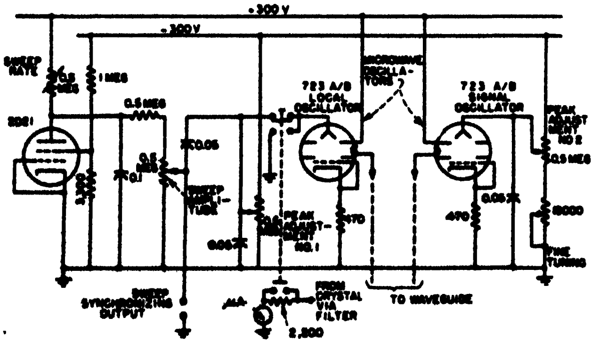
Courtesy Electronics

Fig. 14-37.—Block diagram of Kay "Megasweep."

The sawtooth is generated by a gas-tube relaxation oscillator and serves not only to modulate the local oscillator but also to develop sweep voltage for use in synchronization of an oscilloscope sweep. The power supply furnishes +300 volts, −300 volts (both regulated), and filament power. Ripple is kept very low so as to obtain stable klystron operation.

Sweeping Oscillator

Fig. 14-38 is a simplified schematic diagram of the sweeping oscillator. All adjustments are available at the front panel. The sawtooth sweep-voltage generator uses the miniature 2D21 gas tetrode. The sweep rate control varies the capacitor charging time-constant to make available sweep rates from approximately 50 to 100 cps. The sweep amplitude control adjusts the output frequency excursion up to a maximum of 40 Mc by controlling the magnitude of the sawtooth voltage applied to the local klystron oscillator repeller electrode. The use of this adjustable relaxation oscillator for the production of both frequency and oscilloscope sweep has not only added to the flexibility of the system, but has also eliminated the inconvenient phasing adjustments necessary when the line supply voltage is used for the same purpose.



Courtesy Electronics

Fig. 14-38.—Simplified schematic diagram of sweeping oscillator in Kay "Megasweep."

It is also possible to freqency-modulate the klystron from an external source by removing the 2D21 and feeding the external modulating signal into the instrument through the sweep output jacks.

Mixing Oscillator Circuits

The principal design problem associated with the mixing of the outputs of the two oscillators in the waveguide system involved the efficient isolation of the oscillators from each other to minimize "pulling," or the tendency of one oscillator to drag the other into synchronization with it, when the frequencies are very close. Sufficient decoupling between the oscillators and their load is required. Fig. 14-39 illustrates the waveguide system that was used. The chief feature of this system is the broadband directional coupler, which consists physically of three holes in the wall between the two waveguides.

Directional couplers are designed to provide transmission between two waveguides (or transmission lines) in one direction only. For this particular system, the directional coupler allows energy from the signal oscillator to reach the crystal mixer, but ideally no energy from this source reaches the local oscillator. Similarly, a minimum amount of energy from the local oscillator reaches the signal oscillator. Transmission from one waveguide to the other is directed to the termination and is almost completely absorbed. Thus local and signal oscillators are isolated and the possibility of locking or pulling is minimized though not entirely eliminated, as later explained.

The signal from the signal oscillator undergoes about 10 db attenuation in passing through the directional coupler and, constrained by the one-way properties of the coupler, reaches only the crystal mixer assembly. There is little reflection of this particular signal in either of the waveguides because the crystal assembly is in the shorter waveguide, and a carbon-coated "termination card" is in the other. Both provide loads of approximately the characteristic impedance of their waveguides. The signal from the local oscillator travels directly to the mixer. The portion of the signal from this source, which enters the longer waveguide by way of the three-hole coupler, is directed only toward the termination, where it is completely absorbed.

The two attenuators indicated in Fig. 14-37 have somewhat different purposes. The attenuator between local oscillator and mixer decouples this oscillator from its load. As a matter of fact, the attenuation is inci-

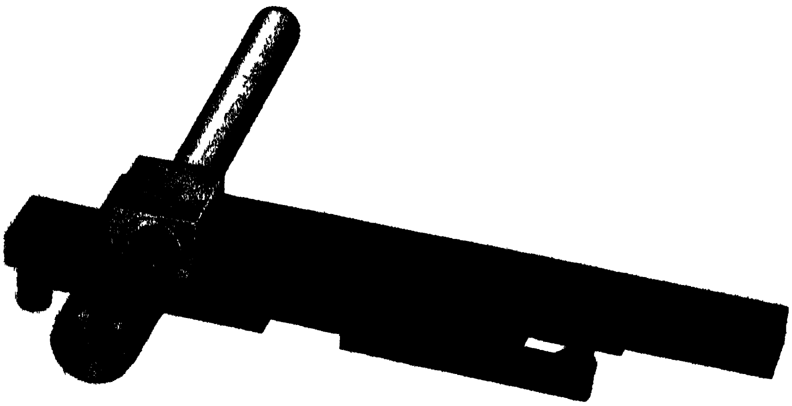


Fig. 14-39.—Precision coaxial absorption wavemeter used in frequency-calibrating circuit.

dental, isolation being the primary function. The attenuator in the signal oscillator path is actually the output attenuator; that is, its adjustment controls the amplitude of the output from the sweeping oscillator. This is possible because the amplitude of a heterodyne output is dependent almost entirely upon the *smaller* of the two components, when the two are of widely differing amplitude. The smaller of the two signals at the mixer is that of the signal oscillator, as this one is attenuated more in passing through the directional coupler. Hence, adjustment of the transmitted amplitude of the signal oscillator controls the output.

This arrangement also prevents amplitude modulation of the local oscillator signal, associated with the frequency-modulating process, from appearing in the output. With the signal oscillator producing an output of constant amplitude, the output of this instrument is, therefore, substantially independent of variations in the amplitude of the local oscillator.

Frequency-Calibrating Circuit

The frequency-calibrating circuit of this instrument is built around a precision coaxial absorption wavemeter, which is shown in Fig. 14-39. The outer conductor of the coaxial line is silver plated on its inner surface, and the movable center conductor, set by a micrometer-type adjustment, is an invar rod successively copper, nickel, and silver plated. This type of construction, together with close mechanical tolerances, insures high Q, negligible temperature coefficient, and high over-all precision.

The crystal current indicated on the microammeter is a function of the rectified output of both the local and signal oscillators. Normally, however, the highly attenuated signal of the signal oscillator contributes a negligible amount to the total crystal current. Hence, in order to monitor the output or measure the frequency of the signal oscillator, it is necessary to stop the local oscillator and increase the meter sensitivity by a pushbutton switch on the front panel.

In the operation of this instrument, a reading proportional to the local-oscillator frequency is taken and recorded, with the sweep amplitude at zero. The pushbutton is then depressed, and a reading is taken of the frequency of the signal oscillator. The output frequency is then proportional to the difference of the readings.

After the oscilloscope pattern is properly located and centered appropriately to the equipment under test, the calibration of the frequency axis of this pattern is accomplished by recording the wavemeter reading proportional to the wavelength of the signal oscillator, which is found by tuning the wavemeter until the amplitude of the entire visible pattern decreases. The wavemeter is then varied until a movable dip is seen on the pattern. The movable dip is an indication of the frequency of the local oscillator and is used to measure the width of the pass-band and, in conjunction with the signal oscillator reading, the absolute frequency. This procedure eliminates the use of any switching.

The midpoint of the frequency excursion can be readily adjusted to a maximum of 500 Mc. This limit is not imposed by the frequency range of the two klystrons. Indeed, if only the range of the two oscillators were considered, a maximum midpoint frequency of nearly 1,000 Mc could be reached. This last figure arises from the fact that each oscillator has a frequency range of about 1,000 Mc, covering the band 8,700 to 9,700 Mc. The upper frequency limitation of the instrument is determined by the properties of the output circuit. Much above 500 Mc, the unavoidable capacitances in the crystal-mixer output circuit take their toll and the output amplitudes fall off. If, however, a coaxial output circuit is used, it is possible to push the frequency limit up to 1,000 Mc.

The sweep amplitude adjustment gives continuous control of the frequency excursion per sweep from zero up to a maximum of 40 Mc. This large frequency excursion, together with the frequency coverage, gives ample and flexible coverage for most if not all commercial vhf applications, including color television.

With the local oscillator sweeping a frequency range that just or nearly includes the constant frequency of the signal oscillator, it is possible to obtain output frequencies close to the audio range. Ideally, of course, the complete audio band could be covered. This is not practically feasible, however, because even with a carefully designed waveguide system, there is some pulling or interaction between the two oscillators when their frequencies are close. This interaction can be ascribed to the small departure of the waveguide terminations and crystal-mixer assembly from the ideal flat waveband loading. It is this pulling which sets a low limit to the frequency range.

In addition to the use of the sweeping oscillator as an aligning instrument, it is also possible to utilize the oscillator without the sweeping feature by simply setting the sweep amplitude at zero. This instrument, so operated, becomes a c-w signal source with the abovementioned ranges.

Markers

No separate marker system is needed. The absorption wavemeter which is part of the equipment is quite satisfactory to produce the characteristic absorption dip in the oscilloscope trace.

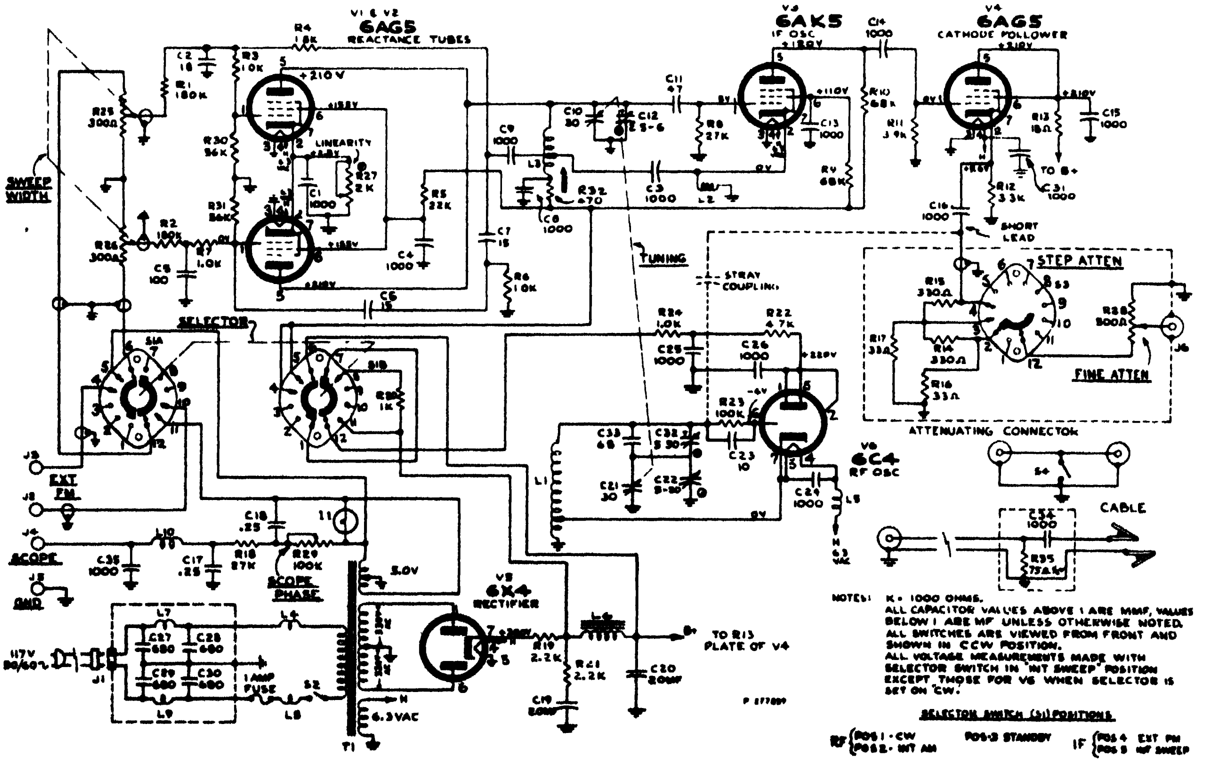
RCA WR-53A F-M Sweep Generator

Although most of the foregoing discussion has concerned signal generators for television receiver servicing — chiefly because this type of equipment has more technically interesting features — some mention should be made of a representative sweep-frequency generator for f-m receivers.

The RCA Type WR-53A is such a piece of equipment. It provides a dependable signal for alignment of the r-f, i-f, and discriminator circuits in f-m receivers.

The i-f oscillator is frequency-modulated by a balanced reactance-tube circuit. The frequency deviation is linear for a frequency swing of ±400 kc at 10.7 Mc, the standard intermediate frequency. The balanced modulating circuit results in maximum oscillator stability consistent with wideband frequency modulation. The instrument contains a special output circuit which results in very low leakage.

The unit comprises a frequency-modulated oscillator for i-f alignment, which is tunable over a range from 8.3 to 10.8 Mc, and an r-f oscillator (c-w or a-m) which can be varied in frequency from 85 to 11 Mc; there is a switch to select the type of operations. The output from the generator is variable from approxi-



NOTES: K = 1000 OHMS.
ALL CAPACITOR VALUES ABOVE 1 ARE MMF, VALUES BELOW 1 ARE MF UNLESS OTHERWISE NOTED.
ALL SWITCHES ARE VIEWED FROM FRONT AND SHOWN IN CCW POSITION.
ALL VOLTAGE MEASUREMENTS MADE WITH SELECTOR SWITCH IN "INT SWEEP" POSITION EXCEPT THOSE FOR V6 WHEN SELECTOR IS SET ON "CW".

SELECTOR SWITCH (S1) POSITIONS

RF {POS 1 - CW
POS 2 - INT AM
POS 3 STANDBY
IF {POS 4 EXT FM
POS 5 INT SWEEP

Fig. 14-40.—Circuit of the RCA WR-53A f-m sweep generator. *Courtesy RCA*

mately 0.1 volt to less than 20 microvolts by means of a "Step and Fine Attenuator." Additional attenuation may be achieved by use of an attenuating connector, an accessory which is connected to the output of the Sweep Generator. It places a low impedance across the output cable, under control of a switch.

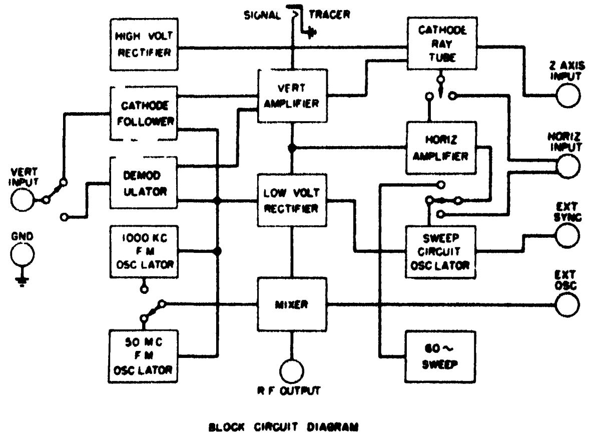
The circuit is shown in Fig. 14-40. The i-f circuit consists essentially of an electron-coupled oscillator fusing a 6AK5, which is frequency-modulated by the 6AG5 reactance-tube modulators. The modulation level, or maximum frequency deviation, is controlled by the "Sweep Width" control *R25-R26*. The modulation frequency is the frequency of the power line or some external source, as selected by a switch.

The output from the i-f oscillator is fed to the 6AG5 cathode follower which acts as a buffer stage between the oscillator and the output circuit. The i-f signal from the cathode follower is fed through an attenuator circuit, previously mentioned, to a jack.

The r-f oscillator uses a Hartley circuit, with a 6C4. The frequency is continuously variable from 85 to 110 Mc. The signal is fed by stray capacitance coupling into the attenuator circuit, from which an output from 0.1 volt to less than 300 microvolts is available at the output jack. By using the attenuating connector, the r-f output range is from 10,000 microvolts to 5 microvolts, approximately. Provision is included for amplitude modulation of the r-f oscillator at twice the supply frequency. Operation of the r-f or i-f oscillator is selected by a switch. Terminals are provided on the front panel for obtaining deflection voltage for a cathode-ray oscilloscope.

Hickok Oscilloscope Model 505A

With progress in the development of radio test apparatus, there appeared to be a demand for equipment in which the cathode-ray oscilloscope and the signal generator were combined in one unit, thereby simplifying operations, and reducing the number of separate test units required, as well as the cost. One such combination is the Hickok Oscilloscope Model 505A, a block diagram of which is shown in Fig. 14-41, and the complete circuit in Fig. 14-42.



Courtesy Hickok

Fig. 14-41.—Block diagram of Hickok 505A cathode-ray oscilloscope containing built-in sweep.

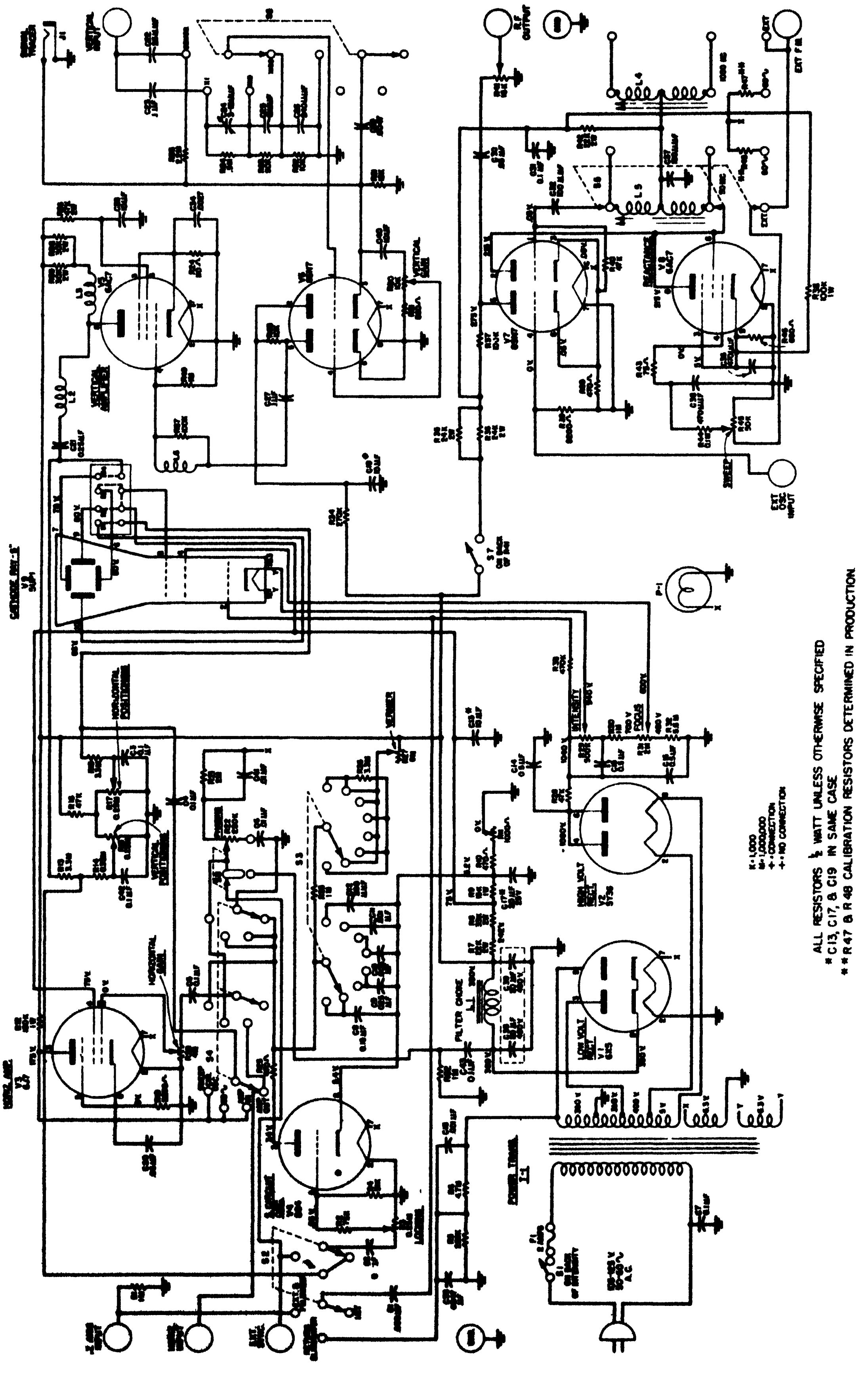
As shown in the lower right portion of Fig. 14-42, the r-f oscillator section comprises a 6SN7 tube *V7* used as an r-f oscillator and mixer, and a 6AC7 tube *V8* used as a reactance tube. The f-m selector switch *S5* determines the frequency of the oscillator and whether it is externally or internally frequency-modulated.

One triode section of the 6SN7 tube is used as the r-f oscillator and operates at a fixed frequency of either 1,000 kc with inductance *L4* in the circuit, or 50 Mc with inductance *L5*.

With *S5* in the 0-30 kc sweep position, the output is 1,000 kc, frequency-modulated at 60 cps with a sweep, variable from zero to 30 kc (15 kc each side), depending upon the position of sweep-control potentiometer *R45*. With *S5* in the 0-450 kc sweep position, the output is 50 Mc, frequency-modulated 60 cps with 0-450 kc sweep, variable, depending upon the sweep-control setting. In either case, the reactance tube *V8* is used as a variable capacitive reactance to frequency-modulate the oscillator. Potentiometer *R41* controls the output level of the oscillator at the r-f output binding post.

The remaining section of *V7* may be used as a mixer, when an external source of r.f. is connected to the "Ext Osc" input, to produce a variable frequency-modulated r-f signal, the frequency range of which will depend upon the frequency range of externally connected signal generator.

The oscilloscope is straightforward, with no unusual features requiring discussion. It shares the power supply with the signal generator. There are all the usual controls, including optional return-trace blanking for 60-cps sweep.



K=1,000
M=1,000,000
+ = CONNECTION
+ = NO CONNECTION

ALL RESISTORS ½ WATT UNLESS OTHERWISE SPECIFIED
* C13, C17, & C19 IN SAME CASE
** R47 & R48 CALIBRATION RESISTORS DETERMINED IN PRODUCTION

Courtesy Hickok

Fig. 14-42.—Circuit of Hickok 505A oscilloscope containing built-in sweep.

CHAPTER 15

TESTING AUDIO-FREQUENCY CIRCUITS

The fundamentals of audio measurements and tests using the cathode-ray oscilloscope apply in the design and maintenance of the audio systems of receivers, public-address equipment, and transmitters.

DEMODULATORS AS AUDIO SOURCES

It is necessary to consider the source of the audio signals and their characteristics. The output of an amplifier is dependent upon the nature of the input signal. If any audio source feeds a distorted signal into an amplifier, the result will be a distorted output. As the demodulator or second-detector stage of a radio receiver is a very common audio source, we may consider some of its features.

Demodulation of A-M Signal

The type of demodulator (detector) found in most receivers produced in recent years is the diode. Its popularity is primarily based on its relative simplicity and low distortion. In the early days of radio, the fidelity and faithfulness of tone provided by the crystal diode were noteworthy, but high-amplitude signals caused it to distort. This fault is corrected in modern receiver designs using the electronic tube diode.

A normal amplitude-modulated signal is shown in Fig. 15-1A. This signal, when applied to a diode detector, results in the audio voltage waveform shown at (B) in the same figure. The output essentially is an undistorted sine wave, showing the voltage across the diode load impedance.

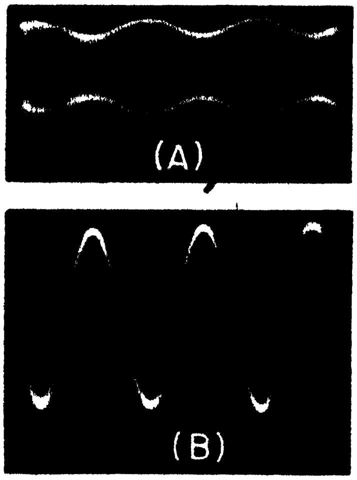


Fig. 15-1.—(A) Normal amplitude-modulated signal; (B) audio output resulting after detection of (A).

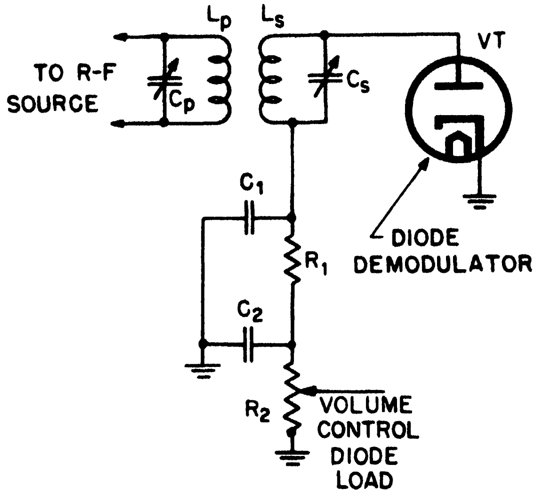


Fig. 15-2.—Standard diode detector circuit. Audio voltage appears across the volume control.

A typical and very common circuit used with diode detectors is shown in Fig. 15-2. *C1*, *R*, and *C2* form an r-f filter. Fig. 15-3 shows the effect of r.f. on the oscilloscope pattern. At (A), there is an indication of the presence of r.f., as shown by a vertical broadness, or fuzziness. At (B) of the figure, the signal has been filtered properly; note that the line appears sharp.

If a circuit parameter changes, such as the diode load impedance, due to the development of a receiver defect, distortion may be caused. A rather common trouble is a variation in an upward direction of the resistance of *R2* in Fig. 15-2. If the proper value is between the

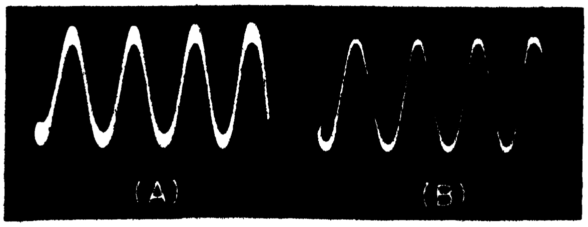


Fig. 15-3.—(A) The broadness of this waveshape indicates that some r.f. is present in the output of the demodulator; (B) the waveform obtained after filtering out the r.f. in (A). Note the differences in the peaks.

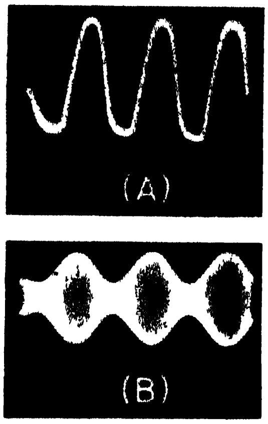


Fig. 15-4.—The top portion of the figure shows the modulated wave from the i-f amplifier. The bottom portion shows the distorted audio voltage output of the demodulator, taken across the diode load resistance.

limits of 250,000 ohms to 1 megohm and the resistance changes to 2 or 3 megohms, considerable distortion results. This is illustrated in Fig. 15-4 where the modulated wave, which is the r-f source signal at the i-f level, is shown; and accompanying it in the lower part of the illustration is the distorted demodulated potential.

The source which drives the detector also has considerable influence on the detector output waveform. An overloaded i-f amplifier waveform output is shown in Fig. 15-5A. The demodulator output for this type of source signal is shown in Fig. 15-5B. Heavy overloading and the corresponding demodulator output are shown in Fig. 15-5C and D. It should be remembered that these are amplitude-modulated r-f signal sources. If the wave is frequency-modulated, nothing significant can be learned by applying the i-f output to the oscilloscope.

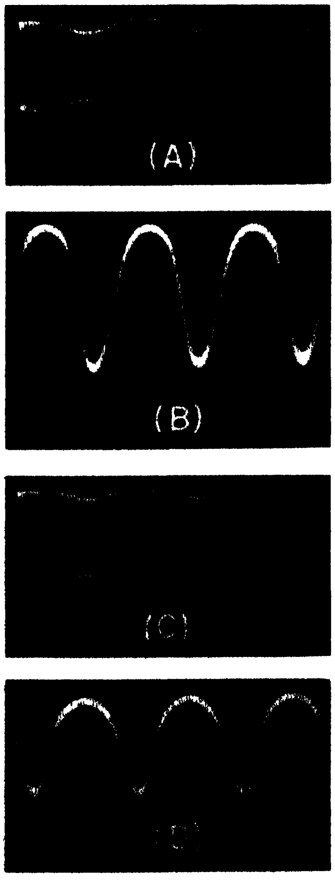


Fig. 15-5.—(A) Waveform resulting from overloaded i-f amplifier; (B) demodulated distorted output for (A); (C) heavily overloaded i-f amplifier output waveform; and (D) increased distortion effect observed at detector output for (C).

If, in the case of high-level modulation, the i-f system in an f-m set is tuned too sharply, amplitude distortion will result and a sine-wave modulating signal will be reproduced as a nonsine wave. If the f-m i-f response curve is not symmetrical, amplitude or phase distortion, or both, may be the result.

Demodulation of F-M Signal[1]

The use of a sine-wave-modulated f-m generator leads to the trace shown in Fig. 15-6A under normal conditions in testing an f-m detector. If there is phase distortion or if a proper balance does not exist, two traces, as shown at (B), will be obtained. Both curves are alike, but at (A) they coincide while at (B) they are slightly separated.

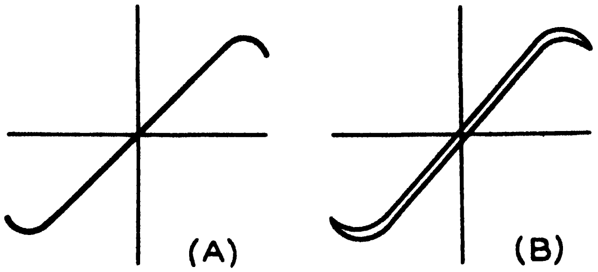


Fig. 15-6.—(A) "S" curve of f-m detector in normal operation; (B) imperfect balance or phase distortion causes a double "S" curve.

The modulating voltage of the signal generator is fed to the oscilloscope external sweep terminals to permit perfect synchronization. If a fault exists in an f-m detector, it is likely to be detected when attempting to align the stage; the nature of the trace obtained on the oscilloscope will be a clue to the cause of the actual trouble.

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[1] Rider, John F., and Uslan, Seymour D., "FM Transmission and Reception," John F. Rider Publisher, Inc., New York, N. Y., 1949.

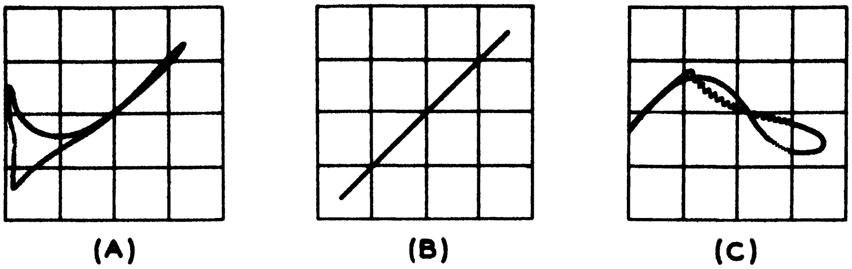

Fig. 15-7. — (A) Pattern obtained when the quadrature circuit of locked-in oscillator is misaligned; (B) correct alignment pattern; (C) pattern obtained when both the quadrature circuit and the last i-f transformer secondary are misaligned.

F-M Locked-in Oscillator[1]

Oscilloscopic patterns of response curves obtained in the visual alignment of a locked-in oscillator are shown in Fig. 15-7. At (A), the quadrature circuit is misaligned. At (B), the circuit is correctly aligned. At (C), the quadrature circuit and the last i-f transformer secondary are both misaligned. When a marker generator is used, the curves appear as in Fig. 15-8. The r-f input to the detector (i-f stage) may be something like that in (A) or (B) of the figure. At (A), the center frequency is identified by the marker. At (B), the lower-frequency limit is indicated. The break at the upper-frequency end is indicated in (C). In any case, the audio signal is to be fed to the input of the audio amplifier.

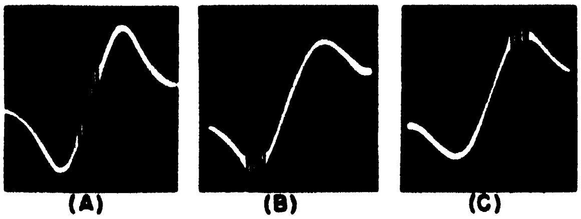

Fig. 15-8.—(A) Marker frequency indication at center of response curve; (B) marker frequency indication at lower limit of the linear portion of the curve; and (C) indication at upper-frequency limit of response curve.

AUDIO SYSTEMS

The oscilloscope is used for studying audio systems as well as the demodulation or detection process. In such a case, it is usually convenient to apply an audio signal of known characteristics, from an appropriate signal generator, rather than to use a received and detected signal. Various types of audio signal generators are available; they may deliver a constant-frequency sine wave, manually adjusted, or a rapidly varying frequency to cover a frequency band, automatically controlled and repeated at a constant "sweep" frequency, or the output may be a square wave rather than sinusoidal; the advantages of the square wave will be discussed later in the chapter. Other special types are available as well.

Basic Test Setup

A simple setup is shown in Fig. 15-9. The signal generator may be any one of several types. The use of a sine-wave generator will be assumed here. First, the waveform of the generator output is checked. A typical sine-wave pattern is shown in Fig. 15-10. The generator is adjusted to some convenient level, such that the first stage of the audio amplifier will not be overloaded. If the amplifier is designed to work with an ordinary phonograph crystal pickup, the input voltage should

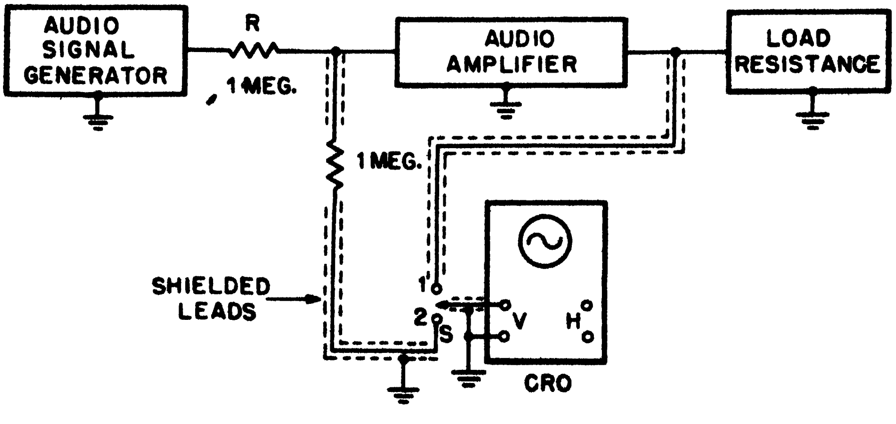

Fig. 15-9.—Simple test equipment setup for checking audio amplifier with signal generator and cathode-ray oscilloscope.

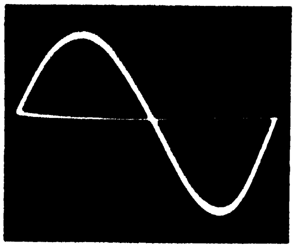

Fig. 15-10.—Sine-wave signal at input to audio amplifier.

not exceed approximately 1 volt. If it is designed for service with a low-level microphone, or phono-pickup of the variable-reluctance type, the output should not exceed 0.1 volt.

Unless the attenuator of the signal generator is properly designed, and used, the amplifier input stage may be overloaded, which will give an erroneous indication of distortion. If the amplifier has a high input impedance, and the generator output impedance is low, the generator will form a shunt on the amplifier input circuit so that the amplifier, when tested, will not be working under normal conditions. To overcome this defect in the test procedure, an isolating resistor may be used, as shown in Fig. 15-9, between the signal generator and the amplifier.

The presence of capacitance in the input impedance of the amplifier may result in a high-frequency voltage drop across *R*. This effect can be corrected by shunting *R* with a capacitor of about 0.001 μf. The adjustment should be made so that there is essentially constant voltage across the input circuit of the amplifier as the frequency of the generator is varied.

If the generator is a simple variable-frequency type, it may be set by hand to a number of convenient test points such as 40, 120, 240, 400, 2,400, 4,000, 8,000 and 10,000 cps to cover the audio range. The height of the trace (sine wave) on the oscilloscope should be relatively constant for a constant output voltage from the generator.

As some generators vary widely in output voltage over the audio band, it is good practice to check the generator output. This may be done by shifting the oscilloscope input directly to the generator output terminals. For many tests, no such precision is required.

When the generator is a sweep sine-wave type, its frequency is automatically varied at, perhaps, 30 times per second. The sweep may extend down to zero and up to as high as 10,000 cps, depending upon the design. Using this form of generator, the external sweep of the oscilloscope is synchronized with the sweep-circuit voltage source in the generator. In this way, an automatic plot of the frequency response curve of the amplifier is secured. In testing audio amplifiers, phase distortion is usually considered unimportant. However, recent developments indicate that phase distortion may be of more importance than was supposed; this subject will be discussed later in this chapter.

There are, then, amplitude, frequency, and phase distortions to be considered. Using the circuit of Fig. 15-9, and assuming the use of a simple sine-wave generator, the waveform of the input and output signals can be compared by the use of switch *S*. It is important that the input and output impedances of the amplifier be like those obtained under working conditions, if worthwhile information is to be obtained from the oscilloscope patterns.

It may be mentioned, however, that the use of a sine wave for distortion testing, with an oscilloscope, is not very effective, since, at ordinary volume levels, the ear will indicate distortion before it becomes apparent on the screen. Therefore, relatively large amounts of distortion may be produced before the condition becomes evident by observation of the pattern. At loud signal levels, however, the oscilloscope is more sensitive as an indicator of distortion than the ear.

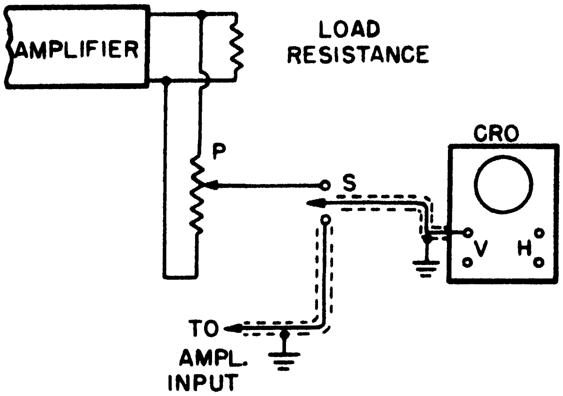

Fig. 15-11.—By controlling the amplifier output with the potentiometer *P* comparison of the input and output waveforms of the audio amplifier is made possible.

Consideration must also be given to the relative difference in signal power levels and voltages at the input and output of an amplifier. If the voltage, for example, is 0.1 volt at the input and 10 volts at the output, the vertical gain control of the scope will need to be reset as the instrument is shifted from input to output. To compensate for this effect, a voltage divider may be placed in the circuit as shown in Fig. 15-11. The value of potentiometer *P* may be about 1 megohm. As the potentiometer resistance is very high, its load-shunting effect is negligible. The arm setting is adjusted so that the height of the trace, with the vertical gain untouched, is approximately the same as that obtained for the input circuit waveform.

Sine-Wave Testing[2]

If a sine-wave input is supplied to an audio transmission system and a distorted-wave output is obtained, the cause of the distortion must obviously lie in the system, which acts on the wave to change its characteristics. The greater the distortion, the more noticeable is the change in waveform.

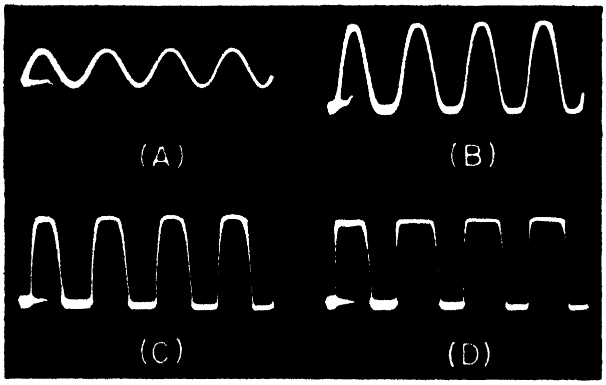

Fig. 15-12.—(A) Sine-wave signal at input to an amplifier; (B) distortion shown by flattening of negative peaks; (C) excessive distortion shown by flattening of negative and positive peaks; and (D) severe distortion indicated by squaring of sine wave.

Fig. 15-12 shows several patterns. At part (A) of the figure, the sine-wave input signal to an amplifier is shown. At (B), (C), and (D), are patterns indicating distortion of the output signals. (B) indicates a flattening of the peaks on negative portions of the signal cycles. Too high a value of grid bias on a voltage- or power-amplifier tube would cause this trouble. Too low a value of plate or screen voltage would cause similar trouble. At (C), flattening of the positive peaks as well as the negative is indicated. Too large an input signal could cause a pattern of this sort to be developed. Increasing the signal input level still further results in greater distortion, as shown at (D). This pattern suggests that a square wave may be obtained by overdriving an amplifier stage.

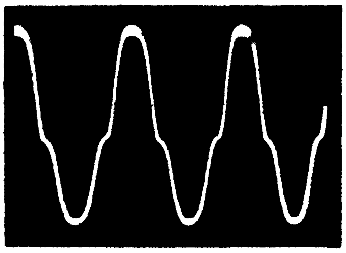

Fig. 15-13. — Excessive bias on the push-pull output tubes of an audio amplifier produced this pattern.

A condition of excessive bias on the push-pull output tubes of an audio amplifier is shown in Fig. 15-13. Too high a value of cathode circuit resistance could be responsible for the excessive bias. The peculiar shape of the waveform is a characteristic of beam power tubes when operated with excessive bias.

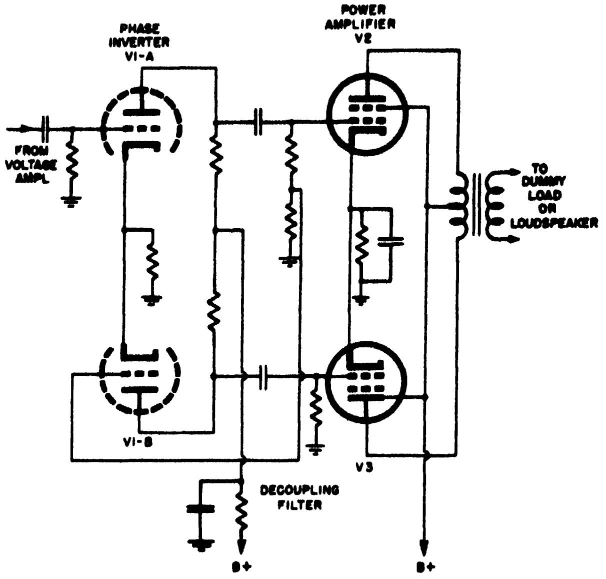

Fig. 15-14.—Audio push-pull power amplifier stage with phase inverter.

A short circuit between the grid and ground in output tube *V3*, in the circuit of Fig. 15-14, produced the pattern shown in Fig. 15-15. If the short had occurred in output tube *V2*, no signal would have been obtained in the output.

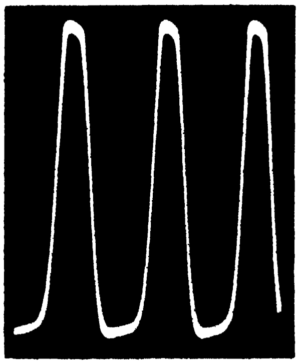

Fig. 15-15.—This pattern was produced with a short circuit between the grid and ground of the *V3* stage in Fig. 15-14.

Fig. 15-16 shows the distortion due to an open grid circuit at one of the push-pull output tubes. The distortion is most evident at the waveform peaks, and indicates a blocking action by the tube, caused by the "floating" grid which does not have a d-c return circuit path to ground.

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[2] Rider, John F., "Installation and Servicing of Low Power Public Address Systems," John F. Rider Publisher, Inc., New York, N. Y., 1948.

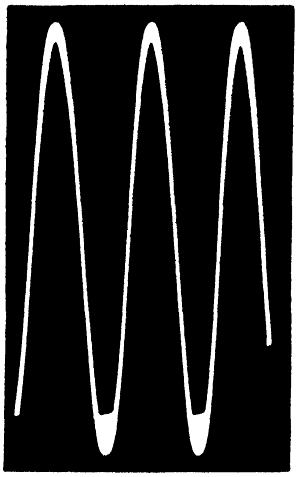


Fig. 15-16.—Distortion due to an open grid circuit at one of the push-pull output tubes.

The pattern shown in Fig. 15-17 illustrates the effect produced by an open cathode bypass capacitor in the output stage. In many output circuits, this bypass capacitor is omitted when the push-pull circuits are balanced. However, if these circuits are not properly balanced due to compromises in design or deficiencies in the output transformer, the bypass capacitor is necessary to prevent undesirable coupling between the two push-pull tubes. In an amplifier of this kind, distortion as indicated will result if this bypass capacitor becomes open-circuited.

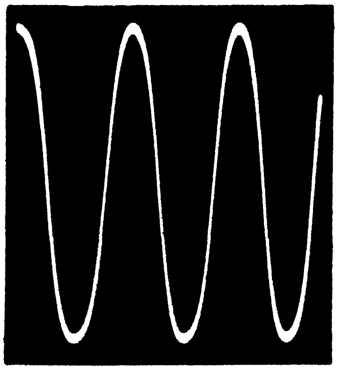


Fig. 15-17.—Effect produced by an open cathode bypass capacitor in the output stage.

Fig. 15-18 illustrates the effect caused by a short in one half of the output transformer primary. A leaky coupling capacitor to one of the output tubes results in the pattern shown in Fig. 15-19. A positive voltage from the plate of the preceding tube is applied through the leaky coupling capacitor to the grid of the output tube. This causes the tube to draw grid current on signal peaks as shown on the waveform.

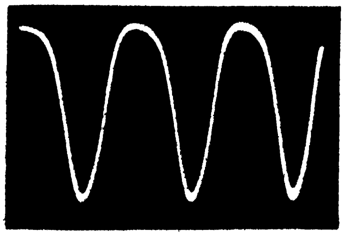


Fig. 15-18.—Effect of a short circuit in one-half of the output transformer primary.

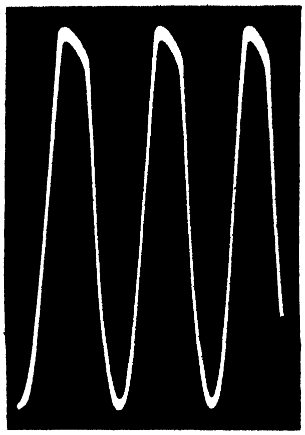


Fig. 15-19.—A leaky coupling capacitor in one of the output tube grid circuits results in this pattern.

One of the advantages of push-pull output over single-ended output stages is that the distortion is greatly reduced. This is because the even harmonics produced by any nonlinearity in the output stage will be cancelled out. If one of the output tubes is not functioning due to tube failure or lack of signal input, these even harmonics will no longer be cancelled out and dis-

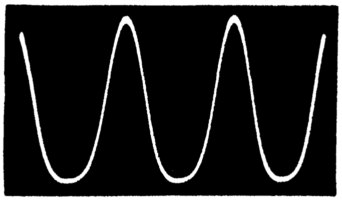


Fig. 15-20 —Distortion due to inoperation of one of the output tubes.

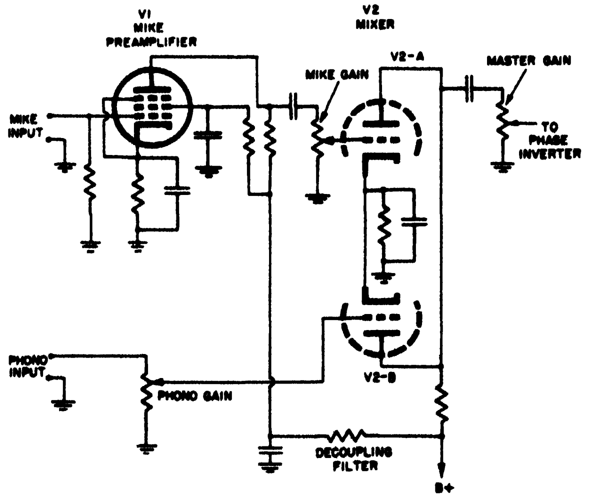


Fig. 15-21.—Mixer circuit of public-address amplifier system.

tortion in the output signal will be apparent. The latter condition is illustrated by the pattern shown in Fig. 15-20. The distortion seen in this waveform is due mainly to the presence of second harmonics which are normally cancelled out by the action of the push-pull output stage.

A mixer circuit of a public-address amplifier system is shown in Fig. 15-21. The pattern resulting from a shorted cathode bypass capacitor in this circuit is shown in Fig. 15-22.

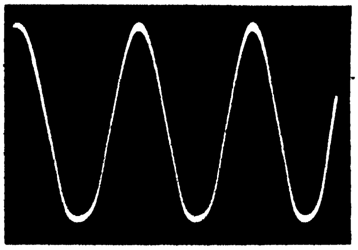


Fig. 15-22. — Pattern resulting from a shorted cathode bypass capacitor in circuit of Fig. 15-21.

Receiver Audio Systems

From the audio system of a receiver, the typical test patterns of Fig. 15-23 were obtained. At a medium volume-control setting, a sine-wave output was obtained, shown at (A) in the figure, indicating negligible distortion. The signal amplitude at the input was equivalent to the average output level of the diode detector of the radio. Advancing the volume control, so that there was some overloading, produced the waveform at (B), showing distortion. By turning the volume control full *ON*, the distortion became more severe, as at (C).

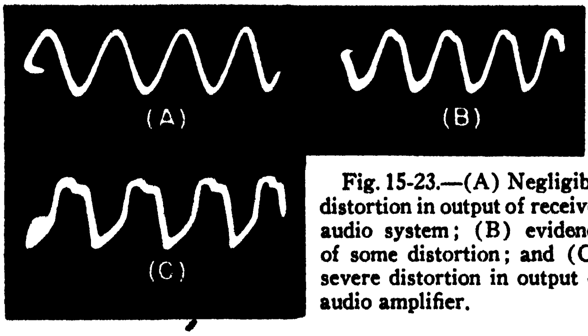


Fig. 15-23.—(A) Negligible distortion in output of receiver audio system; (B) evidence of some distortion; and (C) severe distortion in output of audio amplifier.

Phase Inversion

Many amplifiers use phase-inversion stages to permit push-pull operation. If a nonsinusoidal wave signal is fed to a push-pull stage through a phase inverter that is working properly, the inverter will shift the phase of the distortion. This is illustrated in Fig. 15-24. The signal at one grid of the push-pull stage is distorted on positive peaks; at the other grid, it is distorted on negative peaks, the wave having been shifted in phase by an angle of 180°. It is also quite noticeable that unbalance exists, since the amplitudes are different.

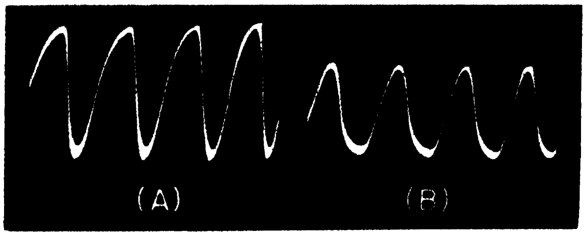


Fig. 15-24.—(A) Distortion on upper peaks of signal in grid input circuit; (B) lower peaks distorted due to phase inversion.

Class B Amplification

The preceding discussion dealt with class A voltage and power amplifier. In a class B system, the conditions are different, and the oscilloscope patterns are correspondingly different. Fig. 15-25A shows the output of a class A driver stage. The driver supplies grid excitation voltages to a class B stage as shown in Fig. 15-26. The coupling transformer *T1* steps down the

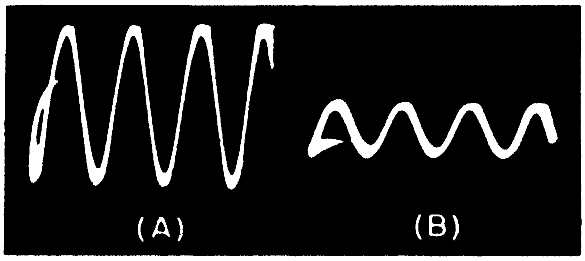


Fig. 15-25.—(A) Output of class A driver stage; (B) input voltage to class B stage.

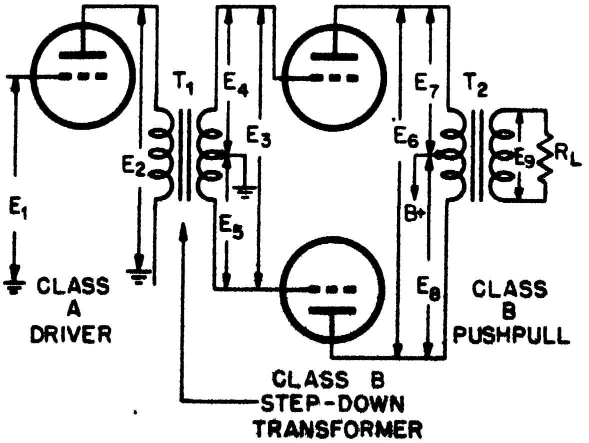


Fig. 15-26.—Circuit of class B output stage with class A driver.

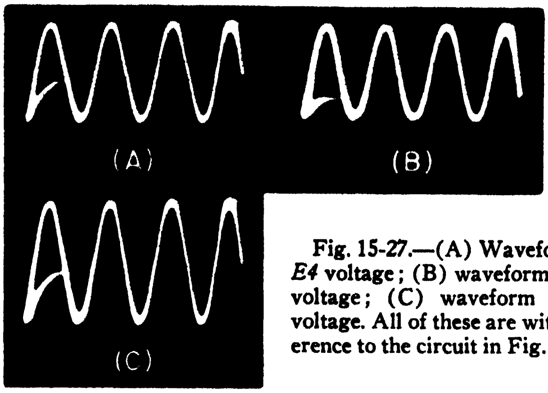


Fig. 15-27.—(A) Waveform of *E4* voltage; (B) waveform of *E5* voltage; (C) waveform of *E6* voltage. All of these are with reference to the circuit in Fig. 15-26.

voltage so that in Fig. 15-25B, the trace is smaller in amplitude than at (A).

The waveforms of the voltages *E*4 and *E*5 are alike and are shown, respectively, in Fig. 15-27A and B. The voltage *E*6 across the output transformer primary is shown at (C) of the same figure. The output voltage across the load resistance R_L is shown in Fig. 15-28A and the voltage at one output tube plate is indicated in Fig. 15-28B. As a class B tube functions only during half a signal cycle, cutoff is indicated. Two such tubes, working on alternate half-cycles, combine to give a full cycle of voltage across the secondary.

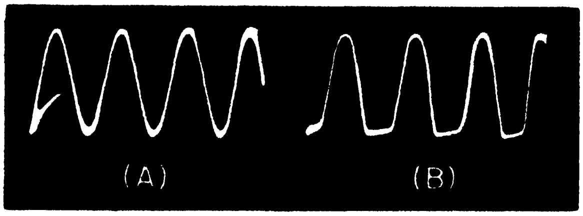


Fig. 15-28.—(A) Output voltage across load resistance R_L; (B) voltage at one output tube plate.

Checking Frequency Response of An Audio Amplifier

The frequency response of an amplifier may be checked using the circuit arrangement of Fig. 15-9. The audio signal generator is a variable-frequency type. Switch *S* is thrown to position *2* to check the input, which should be constant at the various test frequencies. The amplifier gain control may be set at half-maximum since half or less of the full rotation will usually be used in practical applications. The tone control may be set at *Normal*. Frequency runs, if desired, can be made for other tone-control positions.

A one-megohm resistor may be connected in series with the test lead if the shunting effect of the oscilloscope input is noticeable; it usually is not required, and it should not be used unless necessary since it reduces the sensitivity of the oscilloscope.

The height of the trace on the screen may be noted at some reference frequency; 400 or 1,000 cps is commonly used. The output at other test frequencies is then specified with respect to the reference frequency value. The various values may be plotted on graph paper to get the frequency response curve of the amplifier.

Actually, for this kind of work, a db meter or high-impedance a-c voltmeter would be simpler and better than the cathode-ray oscilloscope, but in any case, the taking of such data would be a slow and laborious task. A much more convenient method will be described shortly.

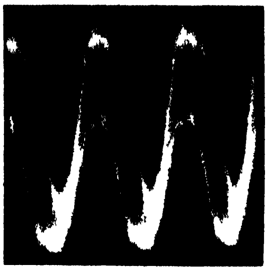


Fig. 15-29.—Oscillation in an audio amplifier.

Oscillation

Fig. 15-29 shows the effect of oscillation in an audio amplifier. If an output voltage waveform of this kind is obtained at the output of a stage but not at the input, the stage is oscillating. A pattern indicating a spurious oscillation is shown in Fig. 15-30. Spurious oscillations occasionally occur in beam-power and pentode output tube circuits. In tracing this type of oscillation, the oscilloscope must be used as the indicator as the frequencies are usually above audibility.

In the work described, a signal generator supplies the signal. In other cases, the receiver audio input, or p-a input, would be driven by a diode detector.

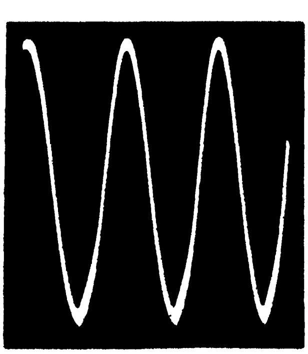


Fig. 15-30.—Pattern illustrating spurious oscillation.

SWEEP-FREQUENCY TESTING

The tedious task of point-by-point determination of the frequency characteristic can be eliminated and an immediate display of the curve presented by a rather simple means. If the audio oscillator supplying the signal is made to vary its frequency, automatically, over the desired range, and at the same time cause the horizontal deflections on oscilloscope screens to represent frequency, then the audio output, applied to the vertical deflection system, will produce a visual characteristic curve. The basic arrangement of such a system is shown in the block diagram of Fig. 15-31.

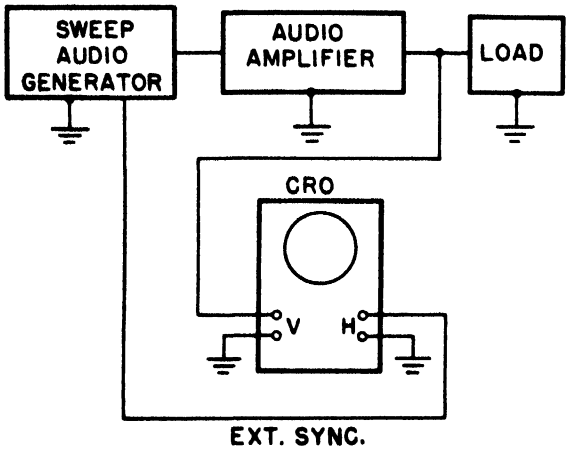


Fig. 15-31.—Test equipment setup for sweep-frequency testing of audio amplifier.

The audio band may be swept continuously from 20 to 20,000 cps, or other frequency bands, according to the design of the generator. The various types of apparatus vary somewhat in details of construction and operation.

Clarkstan Audio Sweep-Frequency Generator

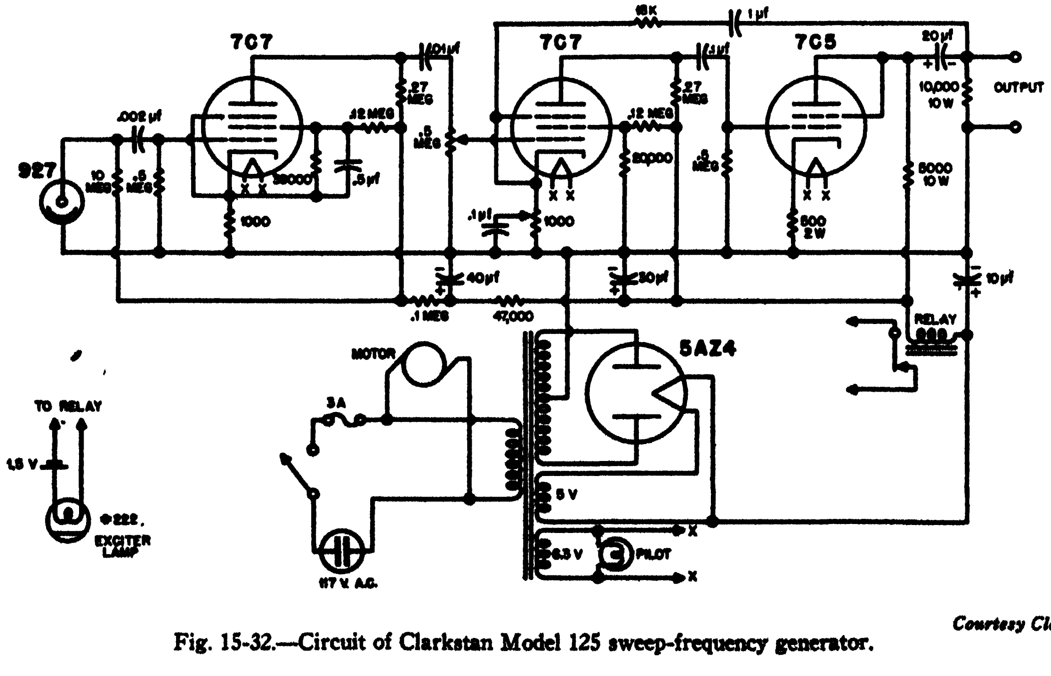
An audio sweep-frequency generator of unusual design is shown schematically in Fig. 15-32. Since the patterns obtained are dependent on the equipment used, a brief discussion of the signal-generating principles involved is in order.

The basic principles of this instrument, the Clarkstan Model No. 125 sweep-frequency generator, are illustrated in Table I. A special disk is used, which has

TABLE I

SPECIFICATIONS
SF Generator Model No. 125

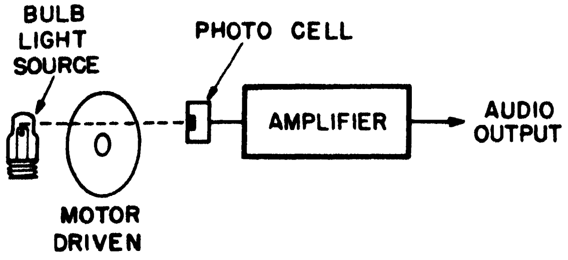
OUTPUT—7 volts, open circuit; 50 milliwatts or 5 volts into 500 ohms.
IMPEDANCE—Internal impedance 200 ohms.
POWER CONSUMPTION—25 watts, 115 volts, 50 and 60 cps.
CABINET MOUNTED—$15\frac{5}{16}$" x 8" x $8\frac{5}{16}$" deep.
WEIGHT—$18\frac{1}{2}$ pounds.
FINISH—Dark maroon Burlene baked enamel.
ALL tubes supplied, 2-7C7; 1-7C5; 1-5AZ4; 1-927.
Will operate with any standard oscilloscope.
Scanning Disk No. 130.1 to be used with SF Generator No. 125
FREQUENCY RANGE—40 cps to 10 kc with 60 cps a.c.
MARKER PULSES at 1, 3, 5, 7, 9, and 10 kc.
SWEEP FREQUENCY governed by 20 synchronizing pulses per second.
(With 50 cps a.c., the range is 33 cps to 8.325 kc with frequency markers at 833, 2500, 4170, 5840, 6250, 7500, and 8325 cps).



Courtesy Clarkstan Corp.

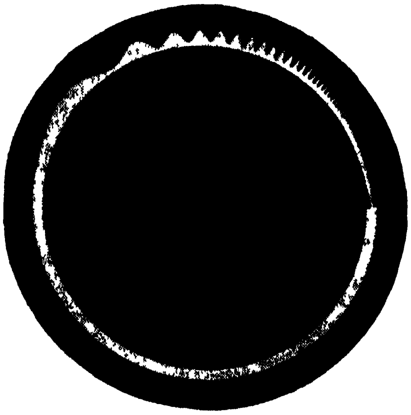
Fig. 15-32.—Circuit of Clarkstan Model 125 sweep-frequency generator.

the sweep-frequency waveform upon it, as a transparency. Light from the bulb source goes through the special disk, shown in Fig. 15-33, to the photo cell, which then delivers a signal to the input of the audio amplifier. The audio output of the amplifier may then be fed directly to the vertical deflection plates of an oscilloscope, or to the vertical amplifier input of the oscilloscope. The vertical gain control of the oscilloscope and the output control of the sweep-frequency generator are adjusted to convenient settings to permit clear presentation of the patterns and a minimum of overloading and distortion. The oscillograms, Figs. 15-34 through 15-45, depict various conditions that may be obtained. The captions are self-explanatory.



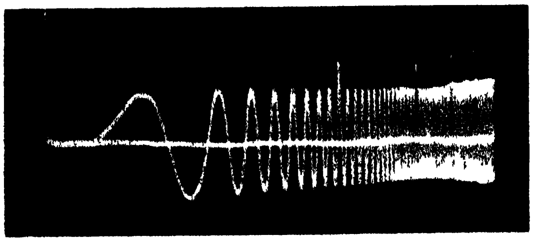
Courtesy Clarkstan Corp

Fig. 15-33 —Basic principle of photoelectric cell, disk-scanning type of sweep-frequency generator is illustrated here.



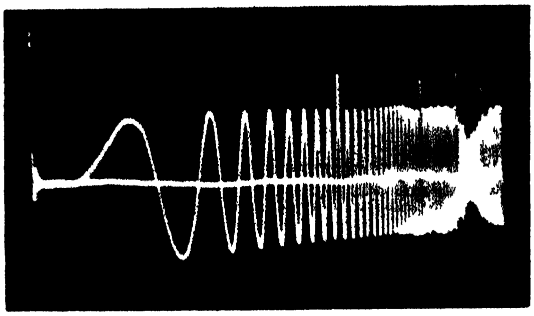
Courtesy Clarkstan Corp

Fig. 15-34.—Card standard for generator system shown in Fig. 15-32. The card is a transparency 5 inches in diameter, scanned with a 0.005-inch × ½-inch light beam. It is rotated at 600 rpm, producing a sweep frequency of from 40 to 10,000 cps, ten times per second, alternating with an unmodulated baseline. The variable width of the baseline compensates for photographic anomalies so that a mean average track density is maintained.



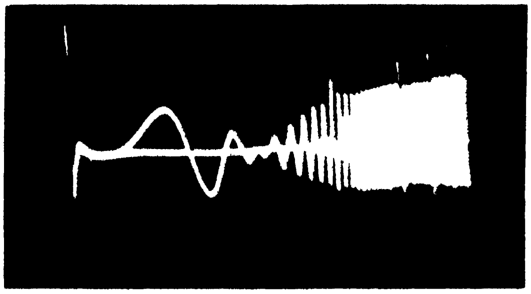
Courtesy Clarkstan Corp.

Fig. 15-35.—The signal, as it comes from the generator, covers a band 40 cps to 10 kc. It is flat over this range to within 1 db. The frequency markers occur at 1, 3, 5, 7, 9, and 10 kc. The baseline is useful for determining relative amplitudes. The reduction of any point to half its amplitude is equal to an attenuation of 6 db.



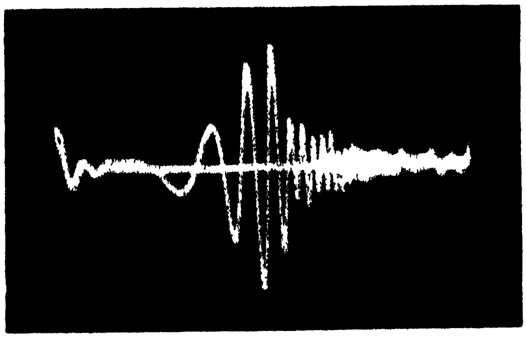
Courtesy Clarkstan Corp

Fig 15-36.—A network composed of *L*, *C*, and *R* across a 500-ohm line. By observation, the resonant frequency is 5,500 cps and the attenuation at that frequency is 6 db.



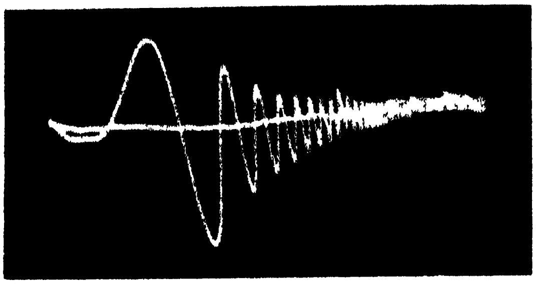
Courtesy Clarkstan Corp

Fig 15-37.—Waveform obtained with a network of *L* and *C* connected across the line. A high *Q* is used. The attenuation is 18 db. The resonant frequency is 450 cps.



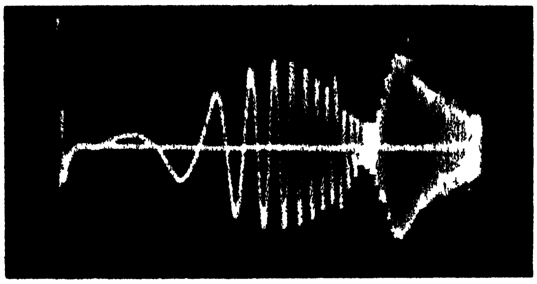
Courtesy Clarkstan Corp.

Fig. 15-38.—Waveform obtained with a band-pass network in series with a 500-ohm line. The network consists of parallel *L* and *C* resonant at 1,500 cps. The system is used to achieve narrow-band operation in communications.



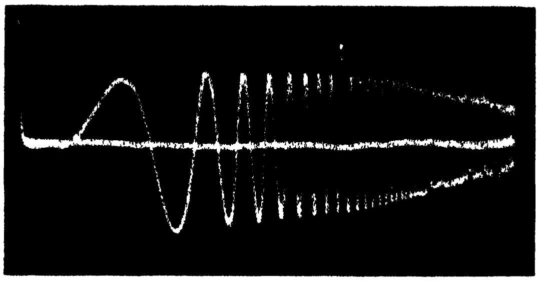
Courtesy Clarkstan Corp

Fig. 15-39.—Pattern obtained using an inductance of several henrys in series with a terminated 500-ohm line. High-frequency transients are attenuated.



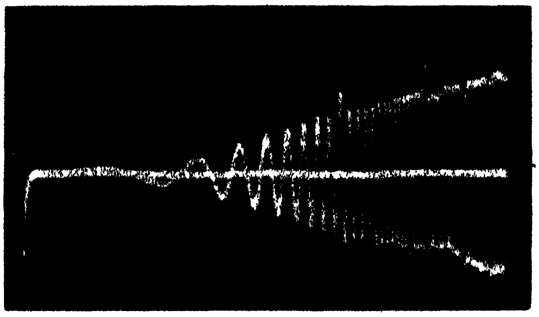
Courtesy Clarkstan Corp

Fig. 15-40.—Effect of two filter networks in the same circuit; one resonant at 500 cps and the other at a high audio frequency.



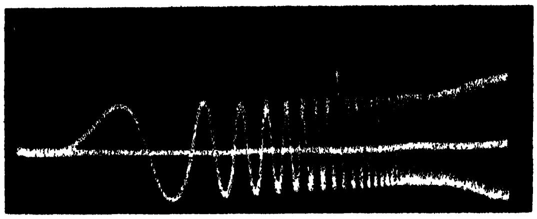
Courtesy Clarkstan Corp.

Fig. 15-41.—Response of an amplifier to 10 kc, showing a gradual attenuation of high-frequency signals, with maximum attenuation of 5 db at 10 kc. An equivalent effect may be produced by a tone control.



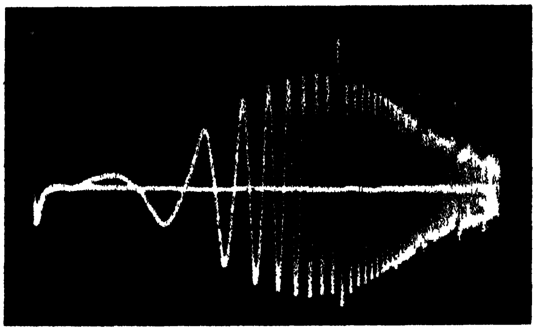
Courtesy Clarkstan Corp.

Fig. 15-42.—Effect of capacitance in series with a 500-ohm line.



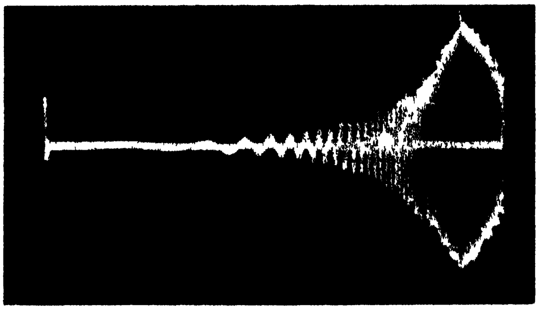
Courtesy Clarkstan Corp.

Fig. 15-43—Response with tone control set for minimum attenuation of high-frequency signals.



Courtesy Clarkstan Corp.

Fig. 15-44.—Pattern obtained in checking response of a magnetically modulated wire in a recording circuit having insufficient bias. Harmonic distortion due to lack of bias and other factors is evidenced by waveform sharpness at low frequencies. A normal 6-db-per-octave attenuation tending toward the lows is due to the velocity effect. From 4,000 cps up, the attenuation is due to the gap effect in which the recording and playback dimensions approach the wavelength on the wire.



Courtesy Clarkstan Corp.

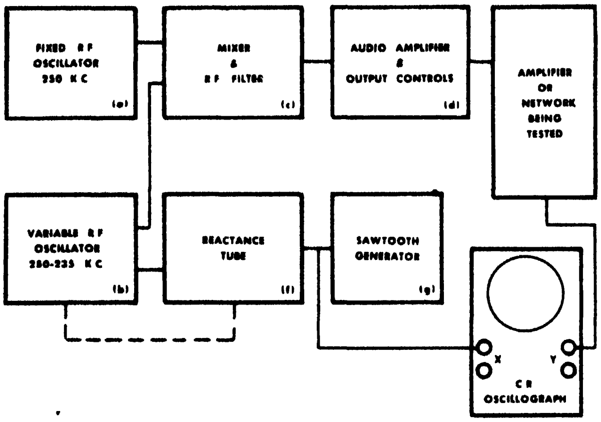
Fig. 15-45.—Effect of a simple *L-C* circuit resonant at 3,000 cps.

Beat-Frequency Audio Sweep Generator

The patterns obtained using an audio sweep generator of the beat-frequency type are somewhat different (for practical rather than theoretical reasons) than those secured with a photoelectric-type generator or mechanically driven unit. It is, therefore, worthwhile to consider some of the fundamentals involved in the generation of an audio sweep using this type of equipment.

As in the previous case, if the frequency of the audio output voltage is varied repetitively over the range sufficiently often, a means is provided of continuously viewing the frequency-amplitude characteristic of an amplifier or other network. Development or adjustment of a circuit may then proceed with the entire performance curve in view, so that it is unnecessary to follow an adjustment at some audio frequency of particular interest and return to the original frequency for observation, at that point, of the effect of a circuit change.

In addition, many communication circuits require that a bridge or inductive balance be maintained over a band of frequencies. Such balances are obtained faster when the effect of each circuit change on the entire frequency band can be made apparent rather than the single effect at a particular frequency. When the audio band is swept less frequently, say once every few seconds, speaker performance may be observed without the inconvenience of "cranking" the frequency dial. Transmission problems may be observed, for initial adjustment, at least, directly on an output meter of normal period.


FIXED R F OSCILLATOR 250 K C
(a)
MIXER & R F FILTER
(c)
AUDIO AMPLIFIER & OUTPUT CONTROLS
(d)
AMPLIFIER OR NETWORK BEING TESTED
VARIABLE R F OSCILLATOR 250-235 K C
(b)
REACTANCE TUBE
(f)
SAWTOOTH GENERATOR
(g)
X
Y
C R OSCILLOGRAPH

Courtesy Clough-Brengle Co.

Fig. 15-46.—Block diagram of a beat-frequency type audio generator.

The block diagram in Fig. 15-46 will aid in understanding the fundamental functioning of the equipment. Sections *a, b, c,* and *d* are the usual stages of the beat-frequency generator in which a fixed r-f oscillator *a* beats with a variable-frequency generator *b* and feeds into a mixer *c* having an audio output that is amplified and controlled at *d* for application to the circuit under test.

The frequency at *b* is controlled by a variable capacitor attached to the frequency-controlling dial. Varying it between 250 kc and 235 kc produces a band of audio frequencies extending from zero to 15 kc. To produce automatic operation, this variable capacitor is replaced by a reactance tube circuit in stage *f*. The reactance tube, in combination with the variable-frequency oscillator, causes the frequency to vary linearly with an applied voltage. With the reactance tube so arranged, it remains only to vary the grid bias of the reactance tube at the desired repetition rate by means of the sawtooth oscillator *g*. The audio output then starts at a low frequency, rises through the audio band to 15 kc and snaps back to the starting frequency to repeat.

When the repetition occurs at a visual persistence rate and the sawtooth wave which is producing the frequency change is fed to the horizontal-deflection plates of an oscilloscope, while the frequency-sweeping signal voltage is fed to the vertical-deflection plates, through the network under test, the frequency characteristic is directly displayed. If any significant change is made in the electrical values of the tested network, the corresponding change in output at all affected frequencies is seen at a glance.

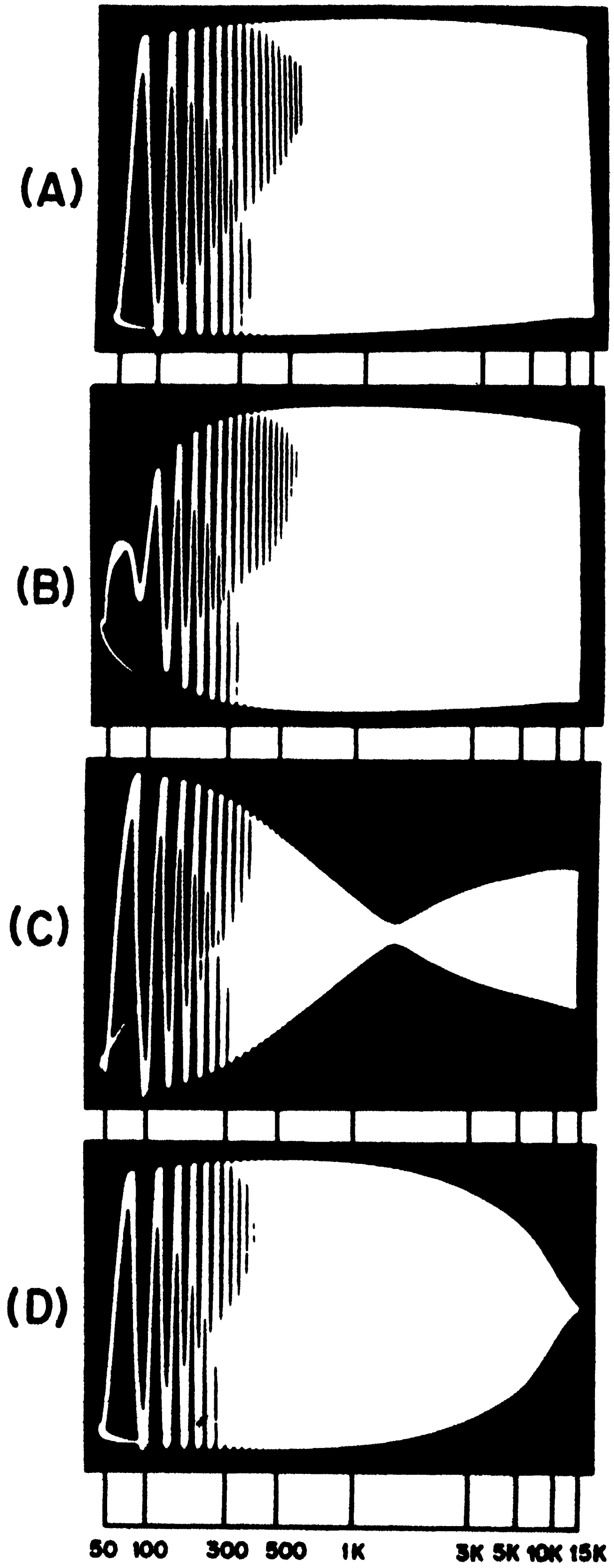
Typical Applications and Circuit Characteristics

Fig. 15-47A shows the basic output trace. (The curved sides of the figure are due to "barrel distortion" in photographing the image on the screen. The actual trace is nearly a perfect rectangle in outline.) At (B), the waveform at 50 Mc is shown, for the case of inserted parallel-*T* null networks using resistance and capacitance. At (C) and (D), the response for null points at 1,500 and 15,000 cps, respectively, is shown.

Fig. 15-48 shows the output of a simple equalizer, with the load resistance high enough to be negligible. (A) applies to a series-capacitor shunt-resistor (that is, low-pass) network, in which $R\omega C = 1$ at 50 cps; (B) is the inverse, a high-pass equalizer, with shunt capacitance and series resistance, such that $R\omega C = 1$ at 15,000 cps. Note that in each case, the amplitude at the frequency mentioned is 0.7 $(= \sqrt{2})$ times the maximum amplitude.

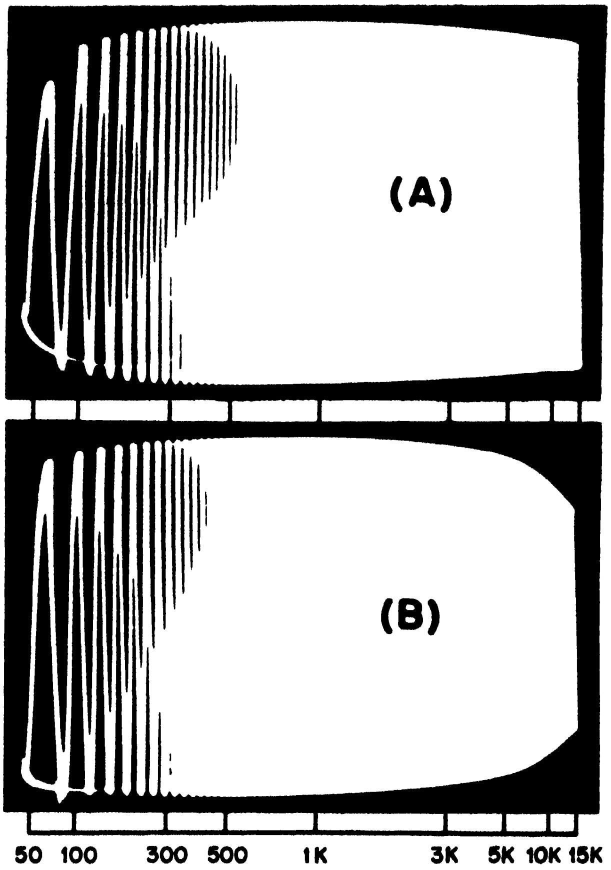
Fig. 15-49 shows the results of balance tests on push-pull transformers. (A) shows a well-balanced transformer; the vertical bar is a calibration line whose amplitude is equivalent to one per cent unbalance. (B) shows serious unbalance at the higher frequencies, caused by stray capacitance.

Fig. 15-50 illustrates the characteristics of carrier phone filters. At (A) is shown a 3-kc low-pass filter (defective in the attenuating band) and at (B) a 4-kc high-pass filter. At (C) and (D) respectively, are shown 4- to 7-kc and 7.5- to 10.5-kc band-pass filter



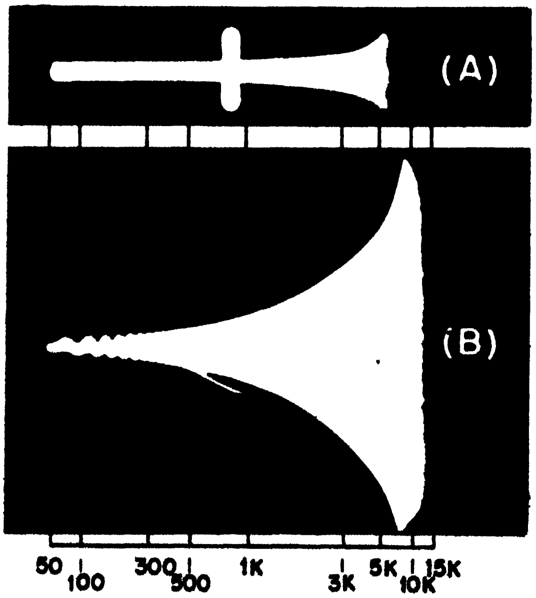
Courtesy Clough-Brengle Co.

Fig. 15-47.—(A) Basic audio output waveform of the beat-frequency generator; (B) response at 50 Mc; (C) at 1.5 kc; and (D) at 15 kc.



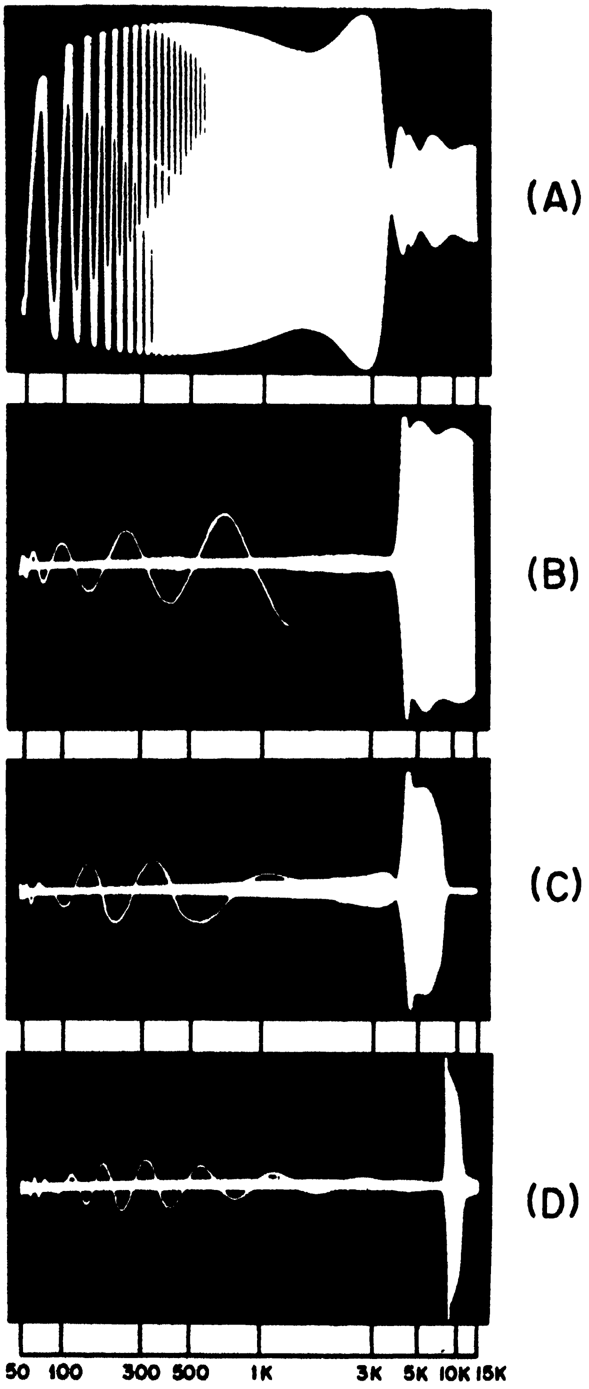
Courtesy Clough-Brengle Co.

Fig. 15-48.—Outputs of simple equalizers, with the load resistance high enough to have a negligible effect: (A) low-pass; (B) high-pass.



Courtesy Clough-Brengle Co.

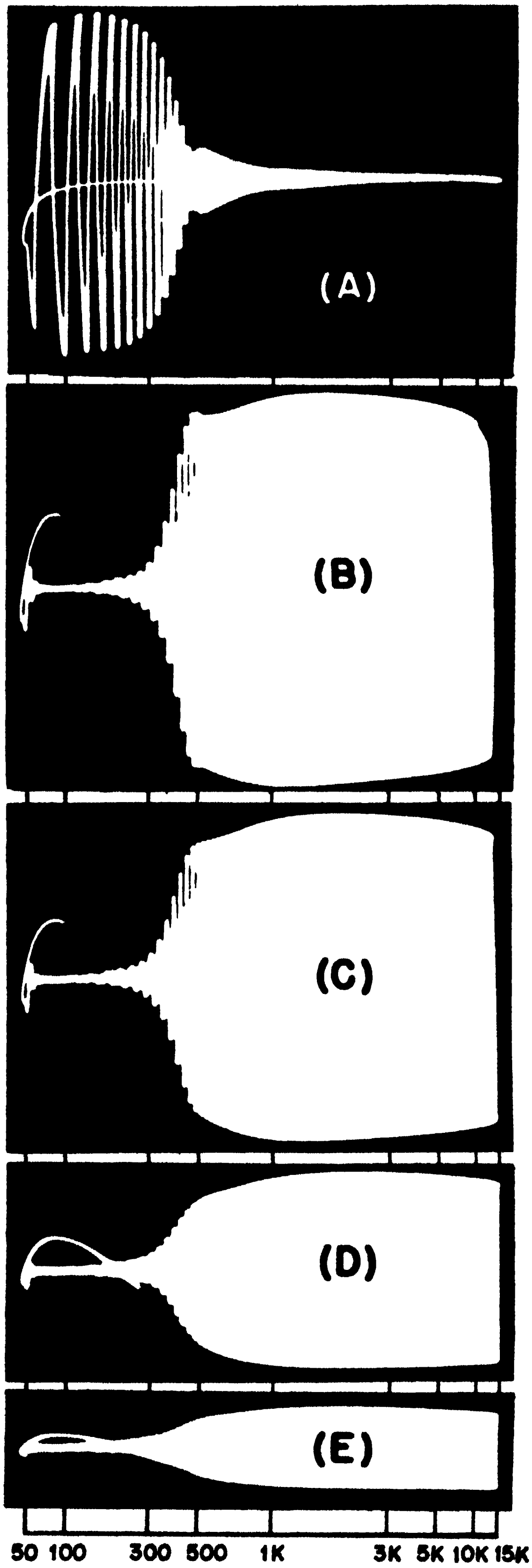
Fig. 15-49.—(A) Result of a balance test on a well-balanced transformer; the vertical bar is a calibration line whose amplitude is equivalent to 1 per cent unbalance; (B) serious unbalance at the higher frequencies, caused by stray capacitance.



Courtesy Clough-Brengle Co.

Fig. 15-50.—Characteristics of carrier phone filters: (A) 3-kc low-pass filter response (defective in the attenuating band); (B) effect of 4-kc high-pass filter; (C) 4- to 7-kc band-pass filter effect; (D) 7.5- to 10.5-kc band-pass filter pattern.

patterns. In Fig. 15-51 are the patterns for a speaker dividing network with 400-cps cross-over. (A) shows the output to the low-frequency speaker. (B), (C), (D), and (E) show the output, at various attenuations, to the high-frequency speaker.



Courtesy Clough-Brengle Co.

Fig. 15-51.—Patterns for a frequency-dividing loudspeaker network, with 400-cps cross-over. (A) Output to low-frequency loudspeaker; (B), (C), (D), and (E), output at various attenuations to the high-frequency speaker.

SQUARE-WAVE TESTING

Sine-wave and sweep-sine-wave tests, while useful, are limited in value and convenience; the use of a square wave, instead of sine, has become very widespread because of its advantages. In preparation for a study of square-wave testing, we may briefly consider some aspects of phase measurements. Information on phase measurements has been given earlier in the book. However, the circuit for checking phase shift by means of Lissajous figures is shown in Fig. 15-52. Various patterns that may be obtained are shown in Figs. 15-53A through 15-53G. The two voltages in question, fed to the vertical and horizontal plates of the oscilloscope so that the deflection amplitudes are equal, produce characteristic patterns which indicate the relative phase.

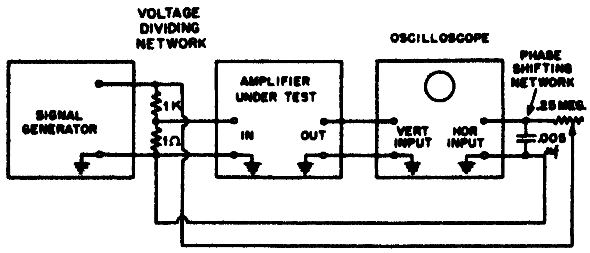


Fig. 15-52.—Block diagram of circuit for checking phase shift by means of Lissajous figures.

Square-wave testing takes into account, in one test, the three factors previously discussed, amplitude, phase, and frequency.

Square-wave testing offers many advantages to the technician. Its relative simplicity and speed make it admirably suitable to quick tests of audio amplifier systems.

Basic Test Setup

The fundamental test setup is shown in Fig. 15-54. In this example, the amplifier load impedance is 500 ohms, a usual value; in any case, the proper value should be used. The oscilloscope is connected across Z_L to indicate the waveform of the voltage E_g across the load impedance.

In testing, feed a square-wave signal of constant voltage into the amplifier input circuit; the oscilloscope may be connected across the signal source to check the input, and then moved stage by stage from the input to the output of the amplifier, checking the effect of each stage on the waveform of the signal. If desired, the over-all effect may be noted by leaving the oscilloscope connected to the output and making various changes or repairs on the individual stages.

The input voltage should be high enough to permit operation of the amplifier, but low enough to avoid overloading of the input stage.

The fundamental frequency of the square wave should be lower or at most equal to the lower-frequency limit of the amplifier, when testing the low-frequency range. A 60-cps fundamental may be used to check the 60- to 6,000-cps range. For the middle and higher frequencies, a 2,000-cps fundamental may be used to check from 2,000 to approximately 20,000 cps.

The effect of a change in the tone-control setting of an amplifier is very marked when using the square-wave test signal. Later, various patterns will be shown to bear this out. The square-wave output will not be

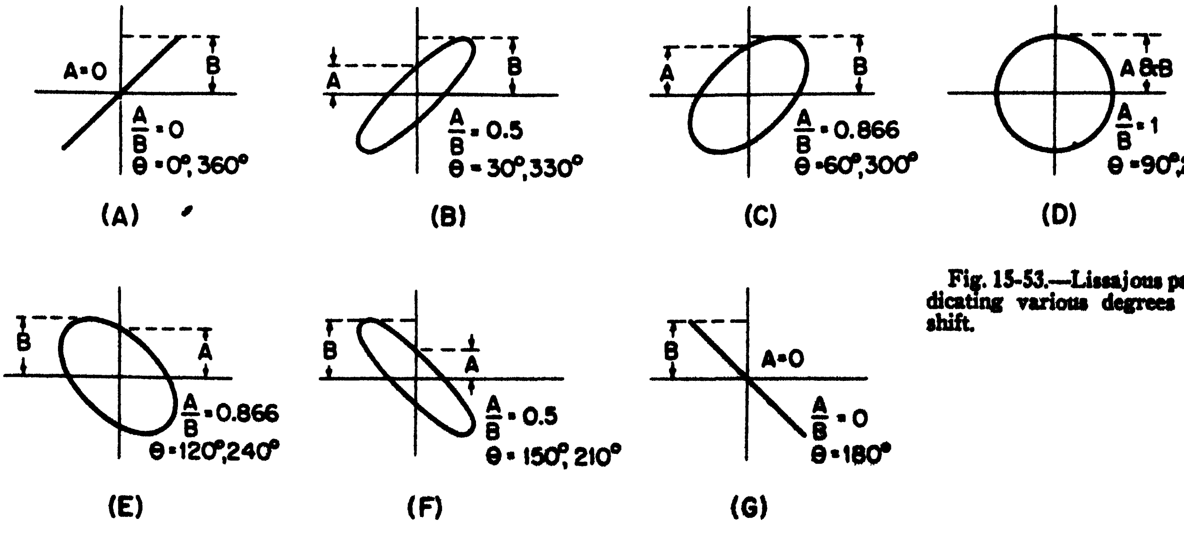


Fig. 15-53.—Lissajous patterns indicating various degrees of phase shift.

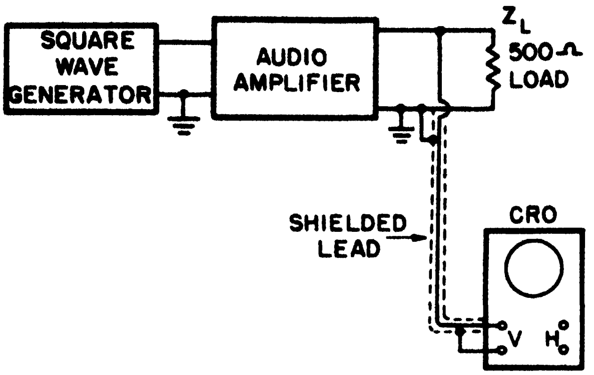


Fig. 15-54.—Basic test setup for checking square-wave response of an audio amplifier.

obtained for relatively low fidelity audio systems used in radio receivers and public-address amplifiers. The departure from a square wave indicates the distortion introduced. A comparatively good amplifier may not reproduce the square wave due to practical commercial considerations which influence the engineering design.

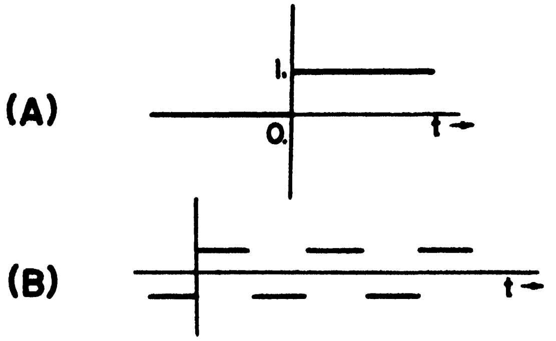
Advantages of Square-Wave Testing

The advantages of square-wave testing are important; the method has become commonly used for several important reasons. It is faster; a complete test is made by measuring at only two frequencies. It is more sensitive to small variations in characteristics; two amplifiers whose sine-wave frequency response is almost identical may have very different square-wave responses. It tells a more complete story; when hum, motorboating, or parasitic oscillations occur they may be rapidly identified and their cause sometimes indicated. The effect of any circuit changes, such as the operation of tone controls, or the adjustment of compensating networks, can be observed instantly without any tedious curve plotting.

Theory

The behavior of an amplifier may be determined by observing its response to a so-called unit function or abruptly steep wavefront, Fig. 15-55A. This visualization is made possible by the cathode-ray oscilloscope. The waveforms in the audio range may be quite complicated; hence it is desirable to use a standardized form of repetition, to get a steady picture that can be checked visually. The square wave is ideal for this purpose.

Any square-wave voltage is simply a succession of positive and negative voltage steps following each other at equal intervals as shown in Fig. 15-55B. If enough time is allowed between steps so that the effect of one step on the output voltage has died away before the next step occurs, the output-voltage change following each step will be independent of the previous steps, and identical to them; this is the reason for requiring a low fundamental frequency for the square wave.



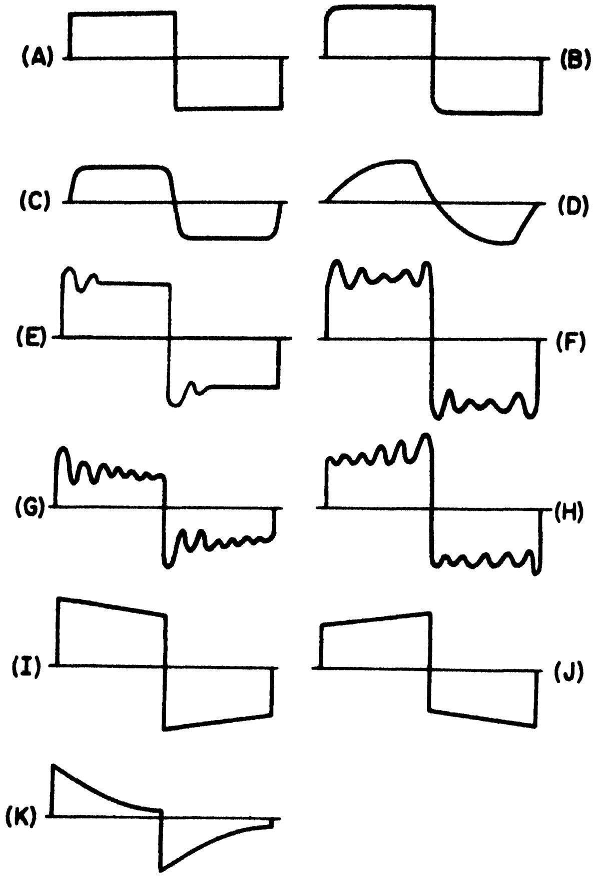
Courtesy Gen. Rad. Exper.

Fig. 15-55.—(A) Unit or step function consisting of a suddenly applied voltage; (B) practical square wave of suitable frequency for test purposes.

The transient characteristic of a circuit is defined as the output which results from a suddenly applied voltage which thereafter remains constant. Knowing this characteristic, the response of a circuit to any type of input voltage can be readily determined. The transient characteristic may be considered to be limited by the high-frequency and the low-frequency transmission of the circuit. The high-frequency transmission will determine the shape of the transient during the interval immediately following the application of the input voltage, while the low-frequency transmission will determine the value of the output voltage after a longer time has elapsed. Obviously an amplifier which transmits down to and including zero cycles (that is, direct current) will eventually reach some steady value of output voltage if the suddenly applied input is not removed, while an amplifier which does not transmit direct current must eventually return to zero output even if the input voltage remains connected.

Wave Analysis

Typical patterns are shown in Fig. 15-56. In general, the time required for the wave to reach its maximum value is dependent upon the frequency range, a rapid rise indicating that high frequencies are attenuated only slightly. The circuit damping is indicated by the shape at the top of the rise; a highly damped circuit will result in a wave which approaches its final maximum



After RCA

Fig. 15-56.—Typical amplifier response patterns: (A) square wave; (B) excess delay (relative phase shift) at the higher frequencies is indicated by a rounding of two diagonally opposite corners of the wave; (C) uniform delay for all the transmitted frequencies is indicated by a rounding of two diagonally opposite corners of the wave; (D) high-frequency attenuation; (E), (F), (G), (H) indicate the presence of oscillations; (I) low-frequency attenuation; (J) low-frequency boost; and (K) poor low- and medium-frequency response.

without any overshoot while under damping will produce a train of oscillations after the initial rise. The natural frequency of the circuit may be determined approximately by multiplying the frequency of the input signal by the number of maxima which occur in one cycle. Excess delay (relative phase shift) at the higher frequencies is indicated by a rounding of two diagonally opposite corners of the wave as shown in Fig. 15-56B. A circuit in which the delay is less than normal at high frequencies will result in changes at the other two corners. Uniform delay for all the transmitted frequencies is indicated by all four corners of the wave having the same shape, as at (C). Certain other characteristics of a circuit may be inferred from its response to square waves. An output waveform containing a train of oscillations (*E, F, G,* and *H*) is always the result of a circuit whose frequency and delay characteristics contain sharp changes. The amplitude and duration of the wave train are measures of the sharpness of the high-frequency cutoff of the circuit, while the period of the oscillation gives a measure of the frequency at which the rapid change in the transmission characteristics occurs. It must be kept in mind that all these effects become more pronounced as the frequency of the square wave approaches the region of poor high-frequency transmission.

If the repetition rate of the square wave is lowered until it falls within the region of poor low-frequency transmission, a slightly different type of interpretation is necessary. The high-frequency effects (rise time, overshoot, and rounding of corners) must be neglected if they are present to a noticeable degree, and only the average shape of the top of the wave is considered. However, most circuits possess a sufficiently wide band of constant transmission that the high-frequency effects are of negligible size when a frequency low enough to show low-frequency deficiencies is used. The low-frequency effects most often found are those due to coupling capacitors, between the plate and the grid of consecutive stages, which cause the transient characteristic to decay exponentially after the initial rise. Typical waveforms obtained with this type of circuit and with a circuit having a rising low-frequency characteristic are shown in Figs. 15-61I, J, and K. No information as to the circuit performance at frequencies lower than the frequency of the square wave can be obtained.

Output waves which are symmetrical with respect to time but whose positive and negative half-cycles are unsymmetrical indicate the presence of nonlinear distortion in the circuit. This effect is usually not noticeable unless oscillations are present in the waveshape. Where the amplitude of the wave trains at top and bottom are unequal, some nonlinear distortion is indicated. Oscillations only on one side of the wave usually denote saturation or cutoff in the equipment under test.

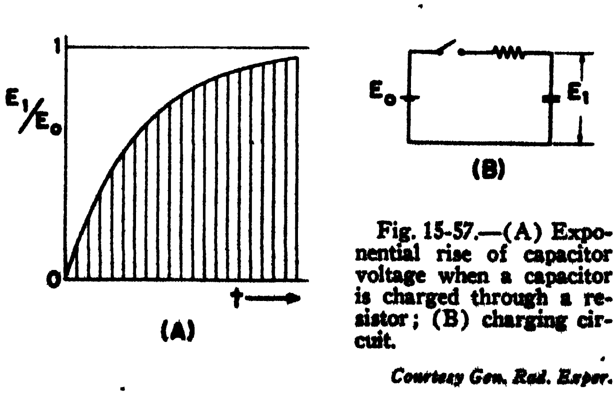
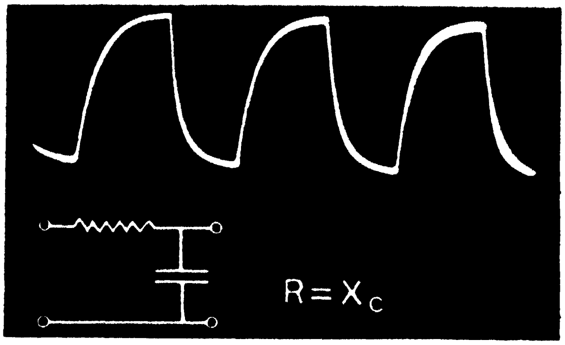


Fig. 15-57.—(A) Exponential rise of capacitor voltage when a capacitor is charged through a resistor; (B) charging circuit.

Courtesy Gen. Rad. Exper.

Square-Wave Response of Simple Circuits

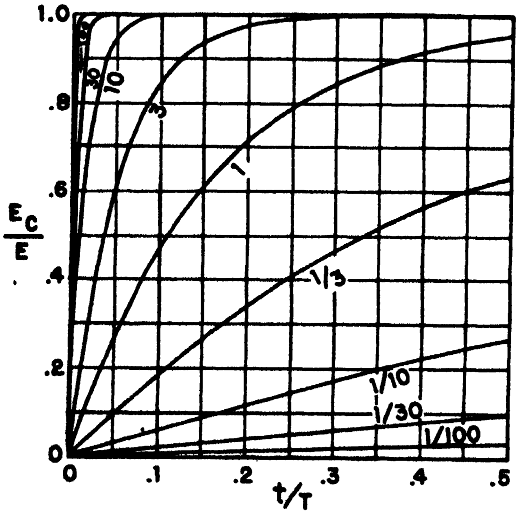
Many of the circuits encountered in practice give very simple and easily recognized transient response curves, which can be used as a guide for other more complicated cases.



Courtesy Hewlett Packard Co.

Fig. 15-58.—A typical oscillogram of the voltage across the capacitor in the circuit shown.

To give an idea of this simplicity, a single case is treated here. Fig. 15-57A shows the familiar exponential rise of the capacitor voltage when a capacitor is charged through a resistor, as in (B) of the figure. A typical oscillogram is shown in Fig. 15-58, and additional information is presented in Fig. 15-59.

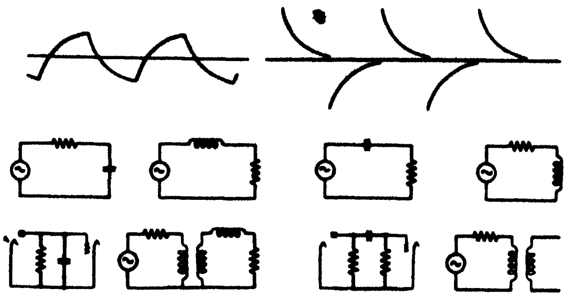


Courtesy Gen. Rad. Exper.

Fig. 15-59.—Curves relating charging time and ratios of source voltage to capacitor voltage.

When a square wave is applied to an amplifier, the output voltage *rises* to a steady value in a way that depends on the *high-frequency* response of the amplifier, and then *falls slowly* to zero in a way that depends on the low-frequency response.

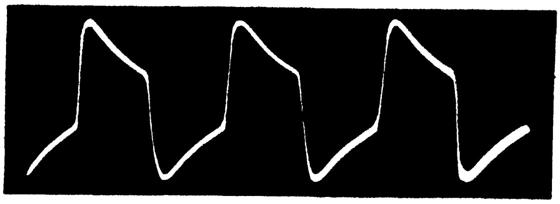
Several such curves are shown in Fig. 15-60. If a square wave is applied to such a circuit, the voltage wave across the capacitor will be as shown at the left in the figure, which is made up of a series of segments. The voltage across the resistor is the difference between the applied voltage and capacitor voltage. This response is shown at the right in the figure.



Courtesy Gen. Rad. Exper.

Fig. 15-60.—Transient response curves and circuits. The waveforms shown correspond to the circuits directly beneath them.

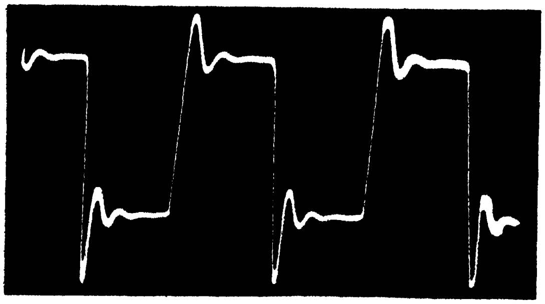
This same series of curves also applies to the case of inductive circuits where capacitance effects are negligible. A number of additional circuits are shown in Fig. 15-60 which shows other practical circuits with square-wave response of similar form.



Courtesy Hewlett Packard Co.

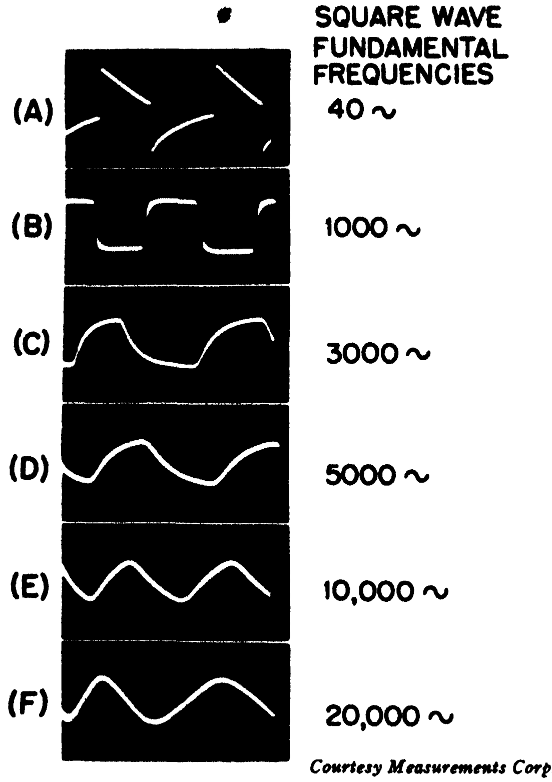
Fig. 15-61.—Result of a square-wave test on a normal amplifier.

In the case of the simple circuits mentioned, it is relatively easy to derive the over-all amplitude-frequency characteristic from the transient response (the converse is not true). In such cases, it would seem quite clear that a knowledge of the transient response is more valuable than a knowledge of the amplitude characteristic, because the latter does not necessarily



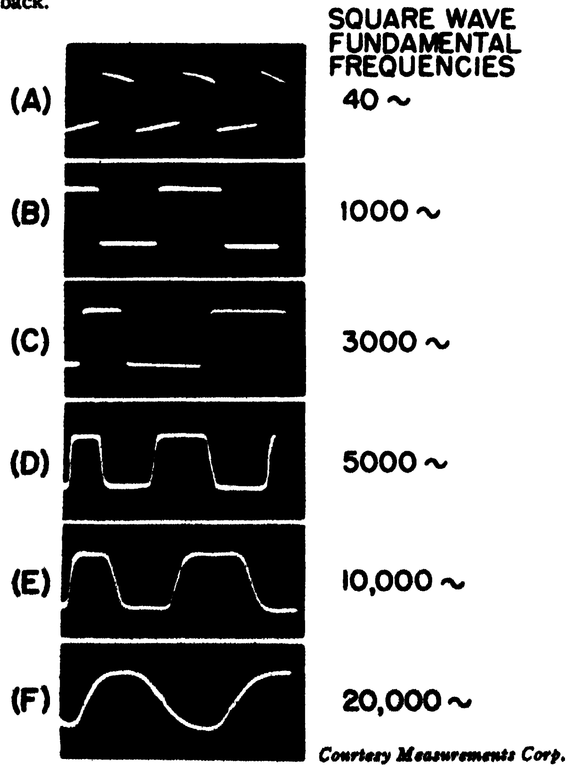
Courtesy Hewlett Packard Co.

Fig. 15-62.—Experimental pattern obtained using negative feedback.



Courtesy Measurements Corp.

Fig. 15-63.—Responses at the indicated square-wave fundamental frequencies of a recording amplifier *without* inverse feedback.



Courtesy Measurements Corp.

Fig. 15-64.—Responses of the same amplifier as Fig. 15-63, but now *with* inverse feedback. Note that here a useful reproduction of the square wave is obtained from about 50 cps to 5 kc, whereas in Fig. 15-63, a tolerable reproduction was obtained only in the region of 1,000 cps.

imply a knowledge of the phase response. In very poor narrow-band amplifiers, it is more difficult to apply square-wave methods because it is difficult to separate the low-frequency cutoff effects from the high-frequency effects. This results in both ends of the characteristic being badly scrambled in a single square wave. In such cases, however, the amplitude response is itself difficult to evaluate in terms of fidelity.

The result of a square-wave test on a normal amplifier is shown in Fig. 15-61. Using negative feedback, the pattern of Fig.15-62 was obtained experimentally. The results of additional tests are shown in Figs. 15-63 and 15-64 in connection with the development of low-distortion recording amplifiers.

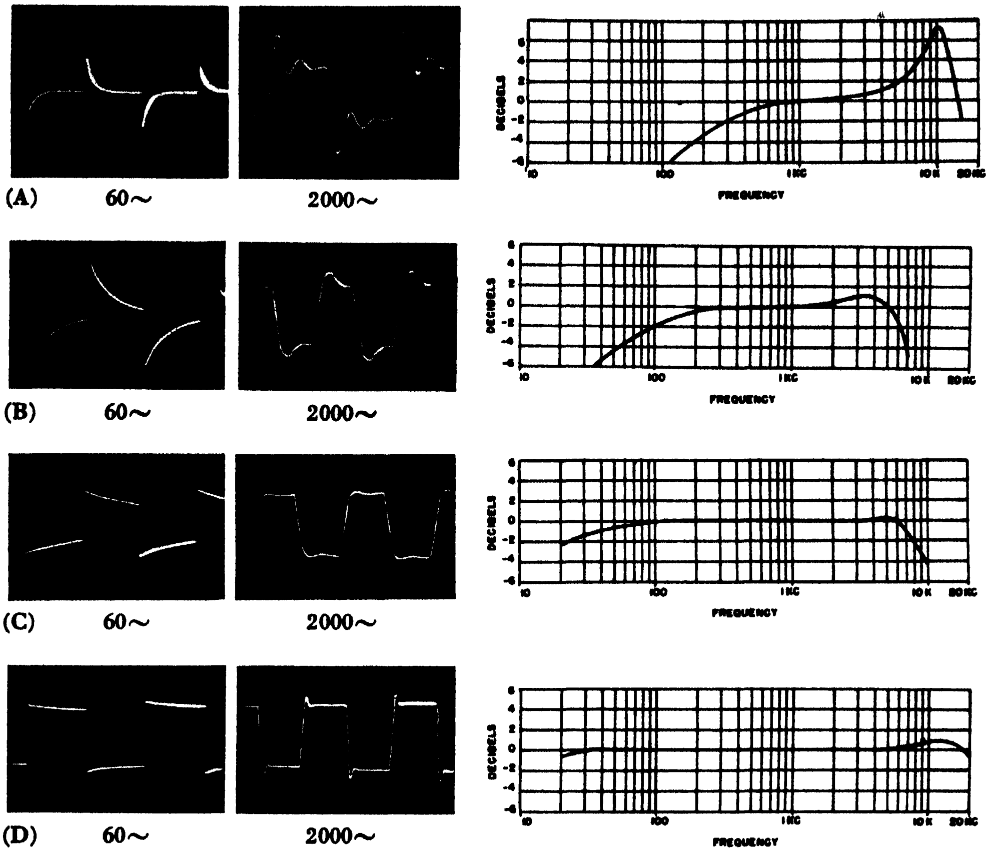
Transformer Tests

The results of some transformer measurements are shown in Fig. 15-65. The frequency response curves correlate satisfactorily with the square-wave test patterns.

At (A) in the figure, the frequency response curve shows a peak at 10,000 cps and also indicates that the low-frequency response curve drops off rather rapidly below 1,000 cps. These conditions are reflected in the square-wave resultant patterns at the left. Using a 60-cps square wave, the rapid decay of the wave shows that the low-frequency response is poor. Using a 2,000-cps fundamental, the rapid rise indicates good high-frequency response and the undulations indicate a resonance effect.

At (B), in the same figure, the pronounced high-frequency peak is absent and there is still appreciable attenuation of the low frequencies. At (C) the frequency response is flatter and the square-wave resultants begin to look more like square waves. The change in waveform, with respect to the low frequencies, is evident more at 60 cps than at 2,000 cps. The slanting downward indicates a loss in low frequencies. The frequency response curve correlates with this information. At (D) an approach to a square-wave output is attained. The presence of a slight peak at the high end of the frequency response curve is indicated by a small pip in the beginning of the 2,000-cps waveform. The slight droop or slant, viewing from left to right, shows a low-frequency loss, in viewing the 60-cps resultant.

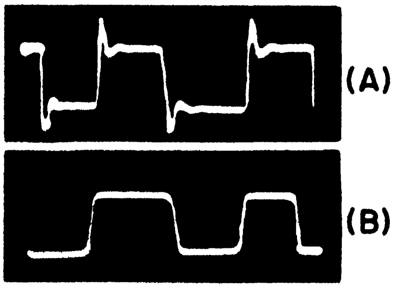
In Fig. 15-66, the waveform obtained using a 5,000-cps square wave, with inverse feedback in the amplifier system, and no load on the output transformer, is indicated at (A). At (B), using the 5,000-cps square wave, with inverse feedback in the amplifier, and a light load on the transformer, the pat-



Courtesy Gen. Rad. Exper.

Fig. 15-65.—Square-wave tests on transformers: (A) frequency response curve shows a peak at 10,000 cps and also indicates a rapid drop below 1,000 cps; (B) absence of pronounced high-frequency peak and evidence of appreciable attenuation of low frequencies; (C) flatter frequency response with some low-frequency attenuation; and (D) approach to a square-wave output.

tern or waveform is seen to improve, indicating the necessity of some loading to obtain linearity and proper amplifier action.



Courtesy Measurements Corp.

Fig. 15-66. — (A) Waveform obtained using a 5,000-cps square wave with inverse feedback, no load on the output transformer; (B) pattern improvement due to loading of the transformer.

Limitations of the Square-Wave Generator

The output waveforms of the square-wave generator all have a certain amount of departure from exactly square, or exactly rectangular. This departure is quite negligible in most cases, since the generator waveforms closely approach square waves. Any possibility of error in qualitative measurements can be eliminated by comparing the output of the equipment under test with the output of the square-wave generator. For quantitative measurements, the rise time of the generated wave should be taken into account.

All the output circuits in the square-wave generator are limited by the internal capacitance and by the capacitance of the external circuits to which they are connected. Because of this, the amplitude and phase of the higher harmonics will depart from proper values for a square wave.

SQUARE-WAVE GENERATION

A square-wave signal for test purposes may be obtained through the use of relatively simple equipment. A simple clipper will be found quite useful for this purpose; the circuit is shown in Fig. 15-67. A signal

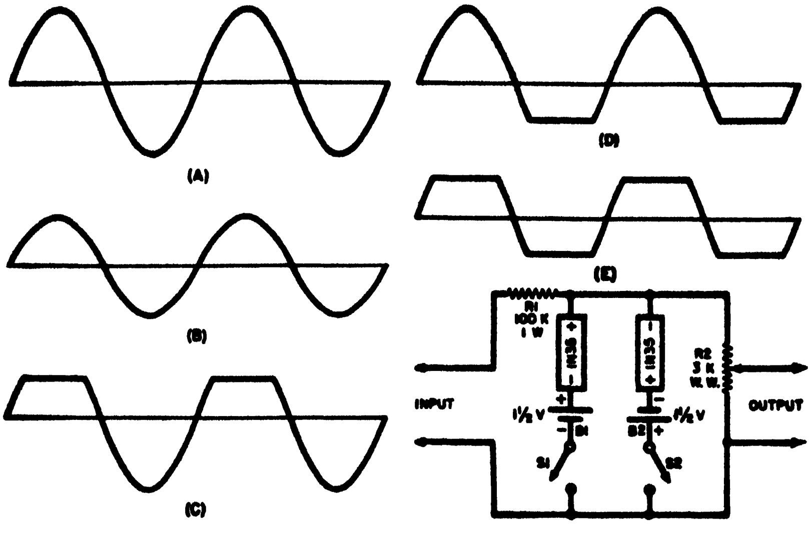


Fig. 15-67. — Complete wiring diagram of crystal diode for varying the waveshape: (A) sine-wave input; (B) sine-wave output, *S1* and *S2* open; (C) pulses essentially negative, *S1* closed, *S2* open; (D) pulses essentially positive, *S1* open, *S2* closed; and (E) output essentially square, *S1* and *S2* closed.

Courtesy Radio News

from a standard sine-wave generator is fed to the clipper input. A true square wave is not obtained, but an approach to it is obtained. The waveshape may be varied as indicated in the caption accompanying the figure.

Three-Tube Diode Clipper

A more elaborate piece of equipment is shown in Fig. 15-68. It has three tubes in the square-wave generator circuit, and one rectifier tube.[8] The instrument includes a triode preamplifier, diode signal rectifier, pentode clipper tube, and self-contained power supply. The preamplifier stage, comprising input capacitor *C*1, gain control *R*1, triode *V*1, and selector switch *S*1, serves to boost the sine-wave input signal. Maximum gain in this stage is approximately 20, and is boosted further by the 2-to-1 interstage transformer *TR*. When the selector switch *S*1 is thrown to the right-hand position, a line-frequency sine-wave signal is picked off the tube heater circuit.

One side of the heater line is grounded and the high side is connected through the switch to the grid-coupling capacitor. When the selector switch is in the left-hand position, the input circuit of the amplifier is connected to the terminals *T*1 and *T*2 to which an external oscillator is connected when square waves of some frequency other than that of the a-c line are required.

*V*2 is a half-wave signal rectifier, a 6H6 tube with its elements connected in parallel. This stage is fed by the secondary voltage of *TR*, which should be approximately 100 volts rms for best results. By means of gain control *R*1, this voltage may be adjusted precisely to the best value. D-c output voltage pulses will be devel-

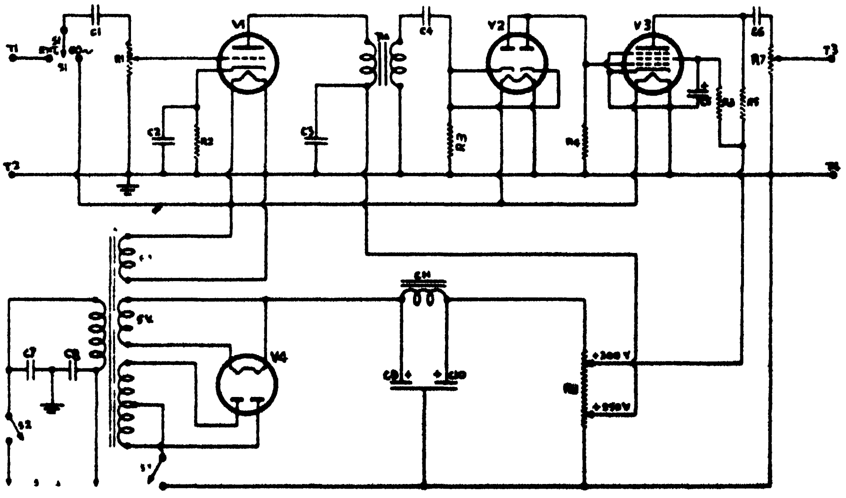


Fig. 15-68. — Three-tube diode clipper circuit.

Courtesy Radio News

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[8] Turner, Rufus P., "Simple Square-Wave Generator," *Radio News*, vol. 31, p. 32, March 1944.

oped across the load resistor *R*4 and applied directly to the grid of the 6SJ7 clipper tube *V*3 each time the lower end of the transformer secondary swings positive.

The cathode of the 6SJ7 is connected directly to ground, and the tube is accordingly unbiased in the no-signal state. The suppressor of this tube is likewise at ground potential, and the screen bypass capacitor *C*5 is returned to the cathode by the shortest possible route. *R*5 is the plate-load resistor for the clipper tube; *R*6, the screen series dropping resistor.

Square-wave output voltage is delivered through the coupling circuit comprising *C*6 (4μ*f*) and *R*7 (100,000 ohms). The latter, being variable, serves as an output gain control. Square-wave output voltage is available at terminals *T*3 and *T*4.

While equipment of the kind described is satisfactory for radio servicing, more exact tests, such as those required in engineering development, may be made with especially well-designed square-wave generators, such as will now be described.

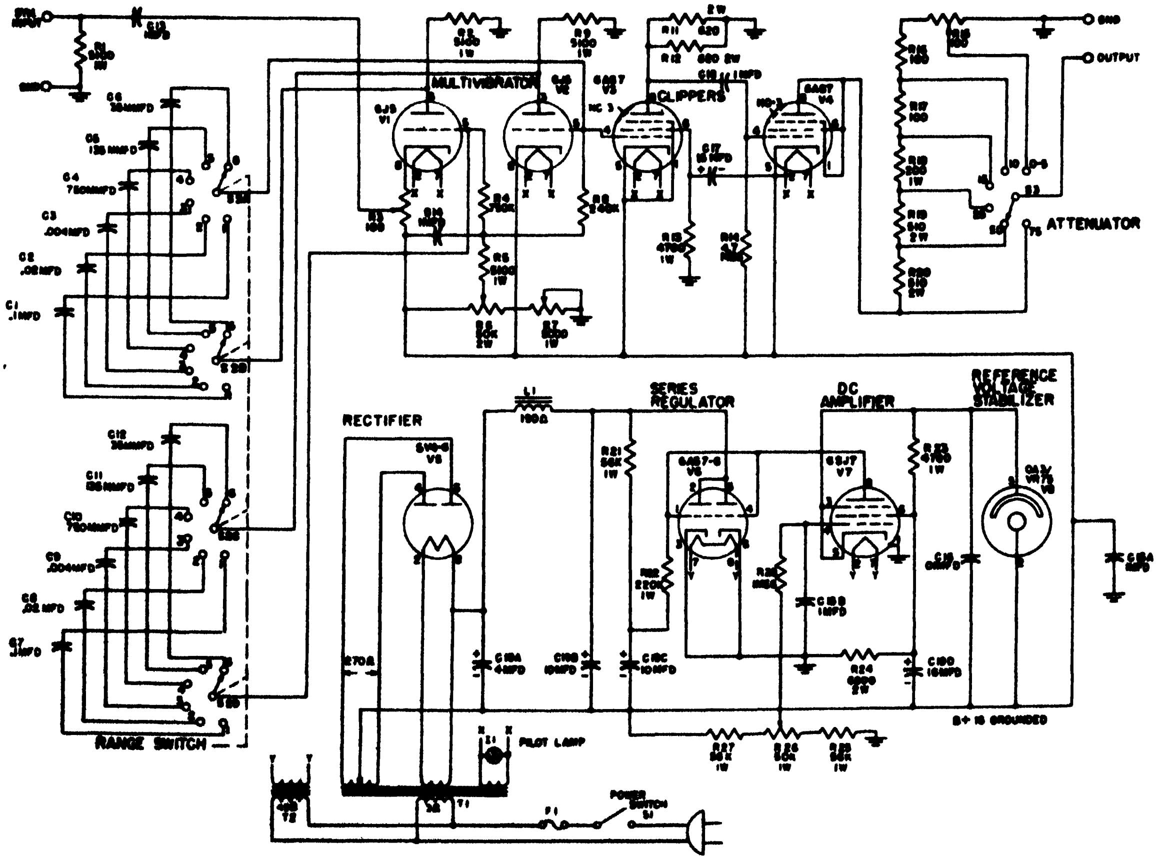
General Electric Square-Wave Generator, Type YGL-1

The General Electric YGL-1, is an excellent square-wave generator. Its circuit is shown in Fig. 15-69.

Multivibrator

The square wave is generated in a multivibrator circuit using two 6J5 tubes. The circuit is, in effect, a two-stage resistance-coupled amplifier, with the output fed back to the input by capacitance coupling. The 6J5 grids are maintained positive with respect to the cathode by means of a variable voltage divider connected across the regulated power supply. A fine adjustment of frequency is made by changing this positive potential.

The frequency of the oscillations is determined by the time taken for the negative grid potential to leak off the grid coupling capacitor, and the time required for the grid to assume a high enough potential for the tube to conduct. Hence, by varying the voltage applied to the grid, this discharge time can be varied.



ALL RESISTORS 1/2W UNLESS MARKED

Courtesy General Electric

Fig. 15-69.—Schematic diagram of GE Square-Wave Generator Model YGL-1.

The coarse adjustment of frequency is accomplished by altering the time constant or discharge time of the grid circuit by changing the value of the grid coupling capacitors, with the frequency-range switch.

The frequency of oscillation can be locked in to an external frequency, by injecting a voltage of this frequency with the cathode circuit of *V*1, the first multivibrator tube. A potentiometer *R*3 permits adjustment of the synchronizing signal, which is applied through a resistance-capacitance circuit *R*1-*C*13.

If the frequency of the multivibrator is a little lower than that of the synchronizing signal, this signal will drive the control grid above the cutoff point a little sooner and the multivibrator will increase its frequency to that of the synchronizing signal. Even without synchronization, the frequency stability is very good, and is satisfactory for many purposes.

Clipper Circuit

The multivibrator is followed by two clipper stages using 6AG7 tubes, which clip off the square wave, thus giving a flat-topped, steep-sided wave output. By tapping off along the plate resistance of the second 6AG7, it is possible to obtain various output voltages, the lowest tap being provided with a potentiometer which allows the output to be reduced almost to zero. The total plate-load resistance of this stage amounts to only 1,500 ohms, since a low value is necessary to obtain the high-frequency response required for steep-sided, square-wave output.

Since the positive side of the d-c voltage is grounded, no output coupling capacitor is used. However, the only d-c voltage that appears at the output terminals is the small drop in the plate-load resistance caused by the plate and screen current drawn by the 6AG7 tube. If the negative side were grounded, an output coupling capacitor would be required to keep the d-c voltage from the output terminals.

Characteristics and Applications

The d-c voltage is supplied by a full-wave regulated rectifier. Because the power supply is regulated, frequency drift of the square-wave output is minimized. A representative figure for drift after warm-up is approximately 0.1 per cent at 10,000 cps over a ten-minute period.

Output

Up to 75 volts is available at the terminals, the instrument having an output impedance of 1,500 ohms at full voltage. The output impedance is approximately 20 ohms per volt, changing with the setting of the output control.

In general, since the impedance of the output rises with an increase in output voltage, care should be taken to minimize capacitances loading the circuit. At 100 kc, with a loading of 50 $\mu\mu f$, good square waves will be produced at 5 to 10 volts output, but at 75 volts, the corners of the waves tend to become rounded.

A suitable coupling capacitor should always be used between the load and the output of the square-wave generator when the load impedance is less than 50,000 ohms. If the load is less than fifteen times the impedance of the attenuator tap being used, there will be a reduction of about 10 per cent in the peak value of the output voltage, but there will be no appreciable distortion of the rectangular wave. A load impedance as low as 50 ohms with a suitable coupling capacitor will also cause a reduction in output voltage, but no distortion of waveform.

Since rectangular-wave testing presupposes adequate oscilloscopic equipment, it is recommended that the input waveshape at the equipment being tested be compared to the output of the generator alone before proceeding with the test work under consideration.

Pulse Width and Shape

The output waveshape of the YGL-1 square-wave generator is a steep-sided rectangular wave with a leading edge rise time of approximately 0.3 microsecond, and a negative pulse width of 25 per cent of the length of 1 cps. Therefore, at any one fundamental frequency, the wave will consist of a positive pulse of 75 per cent and a negative pulse of 25 per cent of 1 cps. The relationship of 75 per cent positive and 25 per cent negative results in a wave of high harmonic content — a feature highly desirable in a square-wave generator.

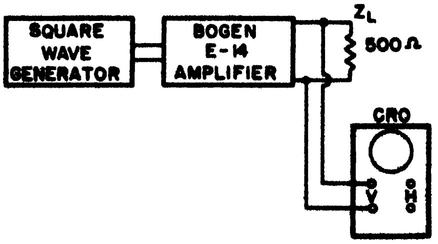
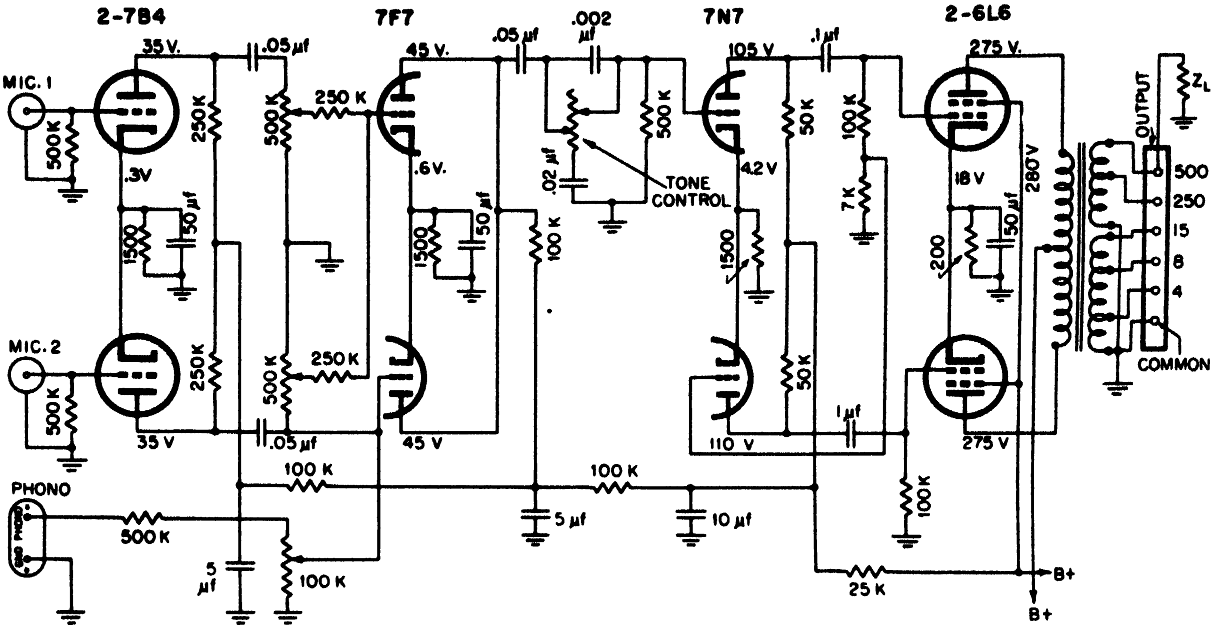


Fig. 15·70.—Test setup for checking the performance of the Bogen E-14 with the cathode-ray oscilloscope.



Courtesy Bogen Co.

Fig. 15-71.—Circuit diagram of public-address amplifier.

LABORATORY TESTS USING SQUARE WAVES

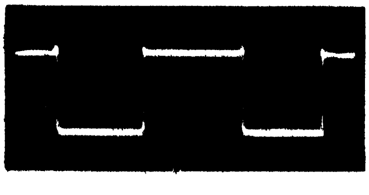
To check theory of testing by square waves in practice, the following patterns were developed and photographed in the John F. Rider Laboratories.

A block diagram of the experimental setup is shown in Fig. 15-70. The amplifier was a standard low-power public-address type having a manufacturer's power output rating of 14 watts; Fig. 15-71 is the schematic of this amplifier.

The input signal voltage was limited to 0.2 volt peak to peak, to prevent overloading the microphone input stage. The amplifier was terminated in its recommended 500-ohm resistive load impedance.

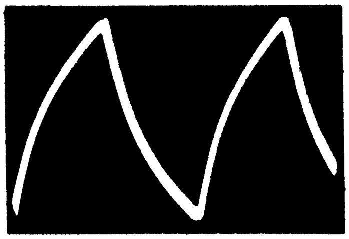
The shape of the square-wave input-voltage waveform, as seen with the oscilloscope connected across the input of the audio amplifier, is indicated in Fig. 15-72. The fundamental square-wave frequency for this series of tests was chosen to be 2,000 cps. Accordingly, the response of the amplifier was checked from 2,000 cps upward to the high-frequency limit of the amplifier.

Fig. 15-72.—Square-wave input to amplifier, having a fundamental frequency of 2,000 cps.



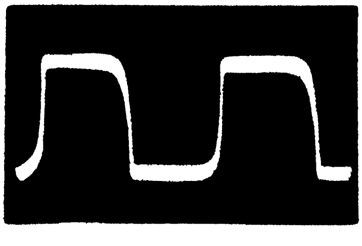
With the microphone gain control on the amplifier set for maximum gain, and the tone control adjusted at *Bass*, the pattern shown in Fig. 15-73 was obtained. The high-frequency response is seen to be extremely poor. This is the voltage waveform across the output load impedance, Z_L.

Fig. 15-73.—Waveform of voltage across load impedance Z_L. Mike gain set at maximum and tone control at maximum bass position (2,000-cps fundamental).



Leaving the microphone gain control at maximum and moving the tone control to *Normal*, the pattern of Fig. 15-74 was obtained. The high-frequency response is seen to be good. The slight downward slant of the top portion of each trace and the rounded right corner are indicative of some phase distortion and low-frequency attenuation.

Fig. 15-74.—Waveform of output voltage across Z_L, with 2,000-cps fundamental signal tone control in normal position and mike gain control at maximum.



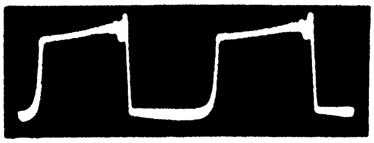


Fig. 15-75.—Waveform of output voltage across Z_L for treble position of tone control and one 6L6 output tube removed; all other adjustments unchanged. Pattern shows slight high-frequency loss, boost in lows and some oscillation or underdamping at the very low frequencies (2,000-cps fundamental).

Changing the tone-control setting to *Treble,* and removing one 6L6, the pattern of Fig. 15-75 was obtained. This shows a slight low-frequency boost, some high-frequency loss, and slight oscillation or underdamping at very low frequencies.

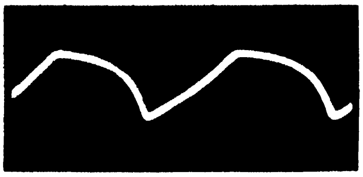


Fig. 15-76.—Voltage output of amplifier, across Z_L, for maximum bass tone control setting and one 6L6 removed from output stage.

Re-installing the 6L6, the "treble" pattern is essentially the same as the *Normal* tone-control pattern, because the 2,000-cps square wave does not check performance at lower frequencies. With the tone control set at *Bass,* to its full extent, and one 6L6 removed, the pattern of Fig. 15-76 was obtained.

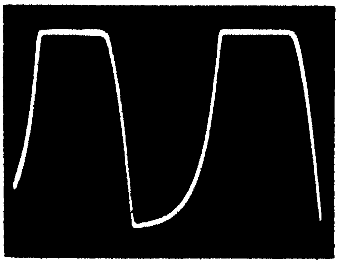


Fig. 15-77.—Waveform of voltage at 7F7 grid with mike gain control at maximum and tone control at bass position.

In Fig. 15-77, the waveform of voltage at the 7F7 grid is shown. The microphone gain was at maximum and the tone control at *Bass* for this pattern. (The tone control, in spite of its position in the circuit, actually does have an effect upon the 7F7 grid circuit.) The voltage waveform at the 7F7 plate is shown in Fig. 15-78. The other conditions are the same as those for Fig. 15-77.

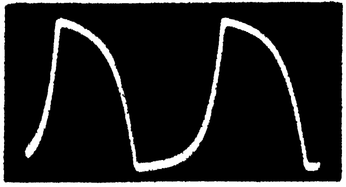


Fig. 15-78.—Waveform of voltage at 7F7 voltage amplifier plate with mike gain control at maximum and tone control in bass position.

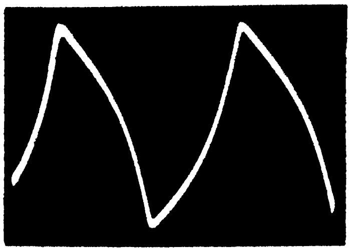


Fig. 15-79.—Waveform of voltage at 7N7 grid with mike gain control at maximum and bass setting of tone control. Frequency and phase characteristics are shown to be definitely affected by the tone control setting.

Continuing the use of the 2,000-cps fundamental frequency, the waveform of Fig. 15-79 is obtained at the 7N7 grid. This wave has sharper peaks than the previous one. Fig. 15-80 shows the voltage wave at the 6L6 grid with the tone control at *Bass* setting. The waveform is unchanged from that at the 7N7 grid. There is delay at the high and low frequencies, low-frequency attenuation, and phase distortion.

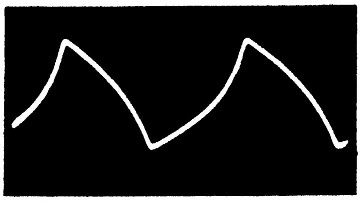


Fig. 15-80.—Voltage wave at 6L6 grid, mike gain control at maximum and tone control in bass position (2,000-cps square-wave fundamental).

The waveform of the signal voltage at the 6L6 grid with the tone control in the *Treble* position is indicated in Fig. 15-81. The distortion is evidently greatly reduced with the tone control adjusted to this setting.

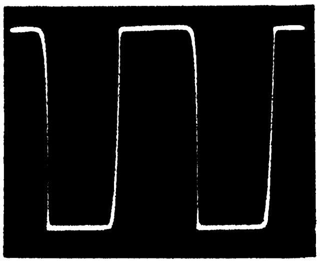


Fig. 15-81.—Waveform of voltage at the 6L6 grid with tone control adjusted to "treble." The distortion is evidently very much reduced at this setting.

Retaining a *Normal* tone-control setting, the signal potential at the 6L6 plate is shown by Fig. 15-82. The thickening of the flat-top portion of the trace indicates the presence of some rectified hum voltage superimposed on the signal. There is some low-frequency attenuation and delay.

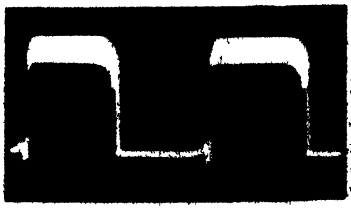
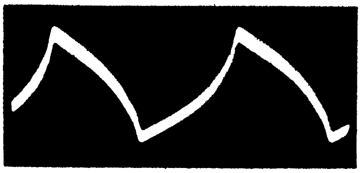


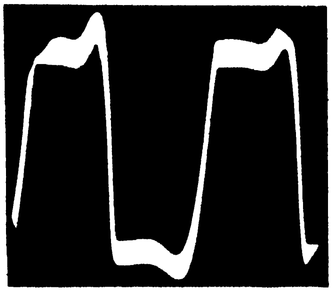
Fig. 15-82.—Voltage at 6L6 plate with superimposed rectified hum component shown in flat-top portions of upper trace; tone control of amplifier at normal setting; (2,000-cps square-wave fundamental).

Fig. 15-83. — Voltage at 6L6 plate with bass setting of tone control. Poor high-frequency response is indicated; also, phase and delay distortion at low frequencies are evident.



Poor high-frequency response, phase delay, and distortion at low frequency are indicated in Fig. 15-83. Tone control was set for maximum *Bass* response.

Fig. 15-84. — Waveform of voltage across Z_L, amplifier output impedance, with output filter capacitor of amplifier open-circuited (2,000-cps fundamental square-wave frequency).



Disconnecting the output filter capacitor, the waveform of Fig. 15-84 is obtained across Z_L, the amplifier output load impedance. The low-frequency peak at the right of the top trace is due to a pronounced hum voltage (low-frequency) output which, in the trace, compensates for the reduction of amplifier gain at low frequencies; this is a result of the deficiency in the filter. The thickened trace is further indication of superimposed hum and is noticed on both halves of the cycle since the 6L6 stage is push-pull.

Fig. 15-85.—Waveform of voltage across output circuit load impedance with cathode circuit resistance of 6L6 push-pull output stage raised to twice the normal operating value.

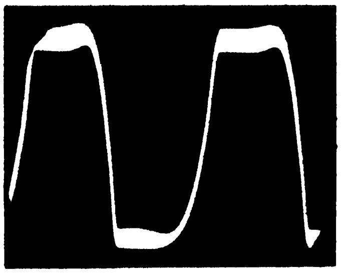


Fig. 15-85 shows the waveform of the amplifier output voltage across Z_L, the load impedance, when a 2,000-cps square wave is fed into the amplifier output circuit and the cathode circuit resistance of the 6L6 push-pull stage is increased to twice the normal value. There is evidently an increase in the hum output voltage and a better approach to linearity and a square-wave output.

Fig. 15-86.—Pattern of amplifier output voltage with plate-to-grid coupling capacitor in voltage amplifier stage short-circuited, mike gain at maximum and tone control set to "normal."

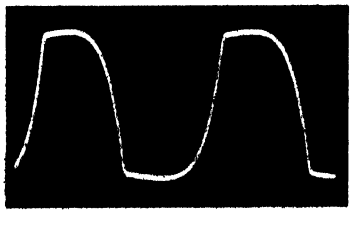
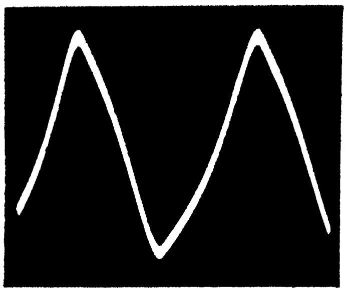


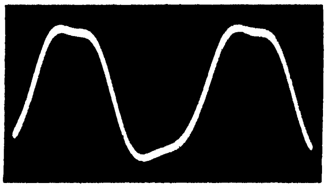
Fig. 15-86 shows the voltage across Z_L, the amplifier output impedance, with a short circuit across the 7B4 plate to 7F7 grid-circuit capacitor, tone control at *Normal* and microphone gain control at full gain.

Fig. 15-87. — 2,000-cps square-wave input to amplifier results in this waveform with tone control set to bass position for maximum attenuation of high-frequency response. Grid-plate coupling capacitor in voltage amplifier stage is short-circuited for this test.



Allowing the short circuit to remain across the capacitor and setting the tone control at *Bass,* the pattern in Fig. 15-87 was obtained. The distortion is evidently much greater. This would be expected in view of the much greater high-frequency attenuation.

Fig 15-88.—Waveform of crystal mike output with square wave fed to amplifier input circuit, mike gain at half maximum and tone control set at bass position; 2,000-cps square wave.



The response in Fig. 15-88 was obtained by feeding a 2,000-cps square wave to the amplifier input and picking up the sound output with a crystal microphone. The experimental arrangement used to obtain this curve is indicated in Fig. 15-89.

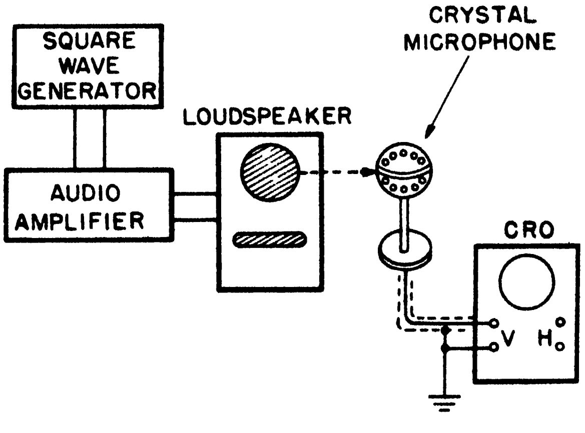


Fig. 15-89.—Setup for checking acoustic output of sound system.

Using a 400-cps square wave with the amplifier tone control on *Bass,* the output of the crystal microphone had the waveform shown in Fig. 15-90. The waveform is evidently quite distorted and the transient response

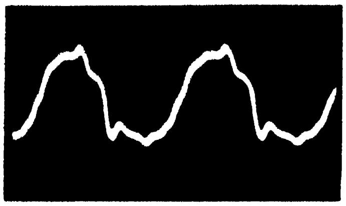


Fig. 15-90.—Waveform of voltage output of crystal mike with 400-cps square wave fed into amplifier input and amplifier connected to a loudspeaker in a bass reflex cabinet. The tone control is set at "bass" position.

of the system is not very good. With the tone control set to *Normal*, the response is even worse, as shown in Fig. 15-91. There is no evidence of square-wave reproduction at all.

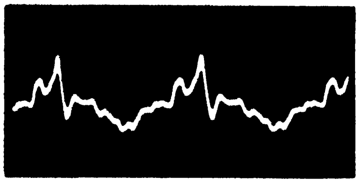


Fig. 15-91. — Waveform of crystal mike output with tone control in "normal" position and using 400-cps fundamental signal at amplifier input. Distortion is very evident.

The voltage output of the crystal microphone has the waveforms shown in Figs. 15-92 and 15-93 for square waves of 800 and 1,000 cps, respectively. The distortion is evidently less, as the frequency increases, since the transient characteristics of the loudspeaker system appeared to be better at the high than at the low frequencies.

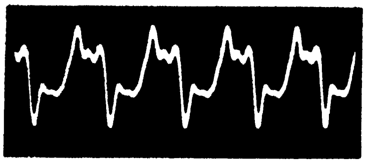


Fig. 15-92. — Crystal mike output when amplifier tone control is set for "normal" response and 800-cps fundamental square-wave frequency is used.

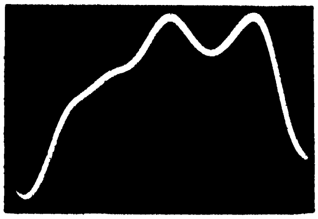


Fig. 15-93. — Waveform of crystal mike output voltage with square-wave fundamental frequency of 1,000 cps and tone control of amplifier set to "normal" position.

The setting of the microphone gain control on the amplifier, its tone control, and the frequency of the square-wave fundamental all affect the pattern. With the gain control set at 50 per cent of full rotation, the following responses were obtained: for 40 cps, Figs. 15-94, 15-95, and 15-96; for 80 cps, Figs. 15-97, 15-98, and 15-99; for 100 cps, Figs. 15-100 through 15-104.

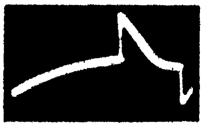


Fig. 15-94.—Tone control of amplifier set at "bass;" fundamental square-wave frequency, 40 cps.

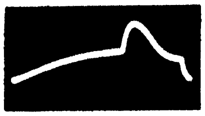


Fig. 15-95.—Tone control of amplifier set at "treble;" fundamental square-wave frequency, 40 cps.

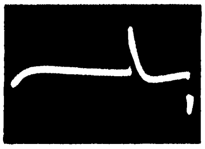


Fig. 15-96.—Tone control at "treble," lower gain control setting (30 per cent of full rotation), 40 cps.

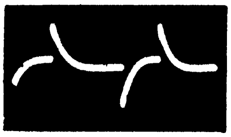


Fig. 15-97. — Test at 80 cps; tone control at "treble."

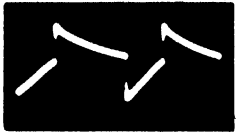


Fig. 15-98.—Tone control at "normal;" 80-cps test frequency.

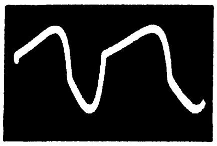


Fig. 15-99.—Tone control at "bass;" 80-cps test frequency.

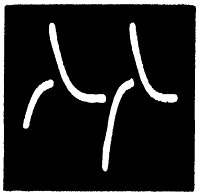


Fig. 15-100.—Tone control at "treble" (50 per cent full-gain volume control setting); 100-cps test frequency.

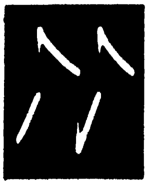


Fig. 15-101.—Tone control at "normal" (50 per cent full-gain volume control setting); 100-cps test frequency.

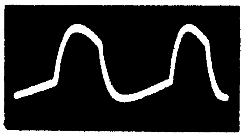


Fig. 15-102.—Tone control at "bass" (50 per cent of full-gain volume control setting); 100-cps test frequency.

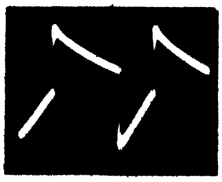


Fig. 15-103. — Tone control at "normal" (30 per cent of full-gain volume control setting); 100-cps test frequency.

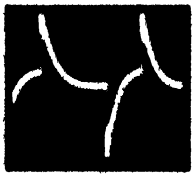


Fig. 15-104. — Tone control at "treble" (30 per cent of full gain volume control setting); 100-cps test frequency.

DYNAMIC MEASUREMENTS ON PEAK-LIMITING AMPLIFIERS

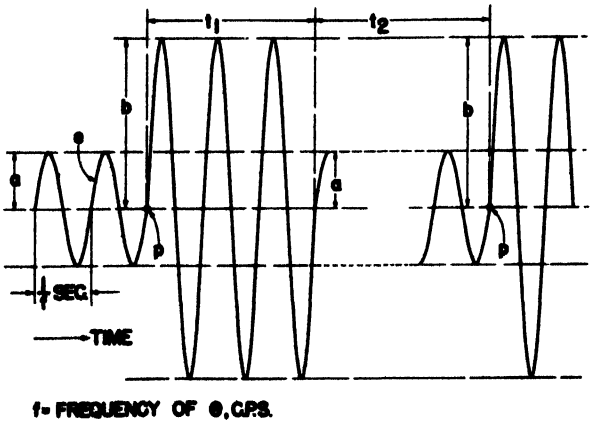
Up to this point, the great value of square waves for amplifier study has been emphasized. There are other methods, however, which are useful, one of which is about to be discussed in the determination of the dynamic characteristics of peak-limiting amplifiers.[4] Peak-limiting amplifiers are used extensively in radio broadcasting to limit overmodulation. Using appropriate equipment, the performance characteristics of such amplifiers may efficiently be evaluated.

The basic technique of the dynamic measurements to be described consists of suddenly applying a sustained signal of high peak amplitude to the input terminals of the peak-limiting amplifier and analyzing the resulting voltage at the output terminals, from the instant the peak signal is applied until the output voltage reaches a steady-state condition.

Technique

The following dynamic-measurement technique is employed. With reference to Fig. 15-105, the technique consists of:

(1) The application of a sinusoidal signal voltage *e* of a predetermined frequency *f* to the input terminals of the peak-limiting amplifier, the amplitude *a* of the signal voltage being of such value that the peak-limiting amplifier is on the threshold of gain-reducing action.



Courtesy Proc. I.R.E. and CBS

Fig. 15-105.—Test signal for evaluating amplifier performance using Donald E. Maxwell technique.

(2) The effectively instantaneous increase of the amplitude of the input signal voltage to a greater predetermined value *b* at a time phase when *e* is crossing the axis of zero amplitude (point *p* of Fig. 15-105) in either the positive or negative direction, as desired.

(3) The maintenance of amplitude *b* of the input signal for a predetermined time interval t_1 and, at the end of that interval, the restoration of the input signal level to amplitude *a* for a predetermined time t_2.

(4) The connection of a cathode-ray oscilloscope to the output terminals of the peak-limiting amplifier, and the initiation and synchronization of a linear-time-base sweep voltage in the oscilloscope, so as to display visually on the cathode-ray tube a predetermined number of cycles of the output signal voltage immediately before and after the application of signal amplitude *b* to the input of the amplifier.

(5) At the end of time t_2, the repetitive re-application for times t_1 of the increased amplitude *b* to the input of the peak-limiting amplifier. Time t_2 may or may not bear a fixed periodic relationship to t_1. It is required only that time $(t_1 + t_2)$ be an integral multiple of $1/f$, and that the sweep voltage of the oscilloscope be so initiated and synchronized that successive traces of the electron beam across the cathode-ray-tube screen are exactly superimposed on one another to produce a stationary visual pattern of the output signal voltage for the predetermined number of cycles before and after the application of the increased amplitude *b* to the input of the peak-limiting amplifier.

(6) The graphical analysis of the amplitude and waveform of the pattern, and comparison of this pattern with applied input voltage to the peak-limiting amplifier.

In a study of peak-limiting-amplifier performance, the time t_1 of Fig. 15-105 is an arbitrary value sufficiently long to allow complete gain reduction of the amplifier to take place. For these specific measurements, the time t_1 was chosen to be approximately 10 milliseconds. It was subsequently found, however, that few if any of the current commercial peak-limiting amplifiers examined reach a stable output amplitude within the 0.01-second observation period, under all conditions of operation.

It is usually required in such an investigation that the time t_2 of Fig. 15-105 be of at least 1 to 3 seconds duration, since the recovery time of most peak-limiting amplifiers is of this order of magnitude. Therefore, the cathode-ray tube used for viewing the transient phenomena should have a long-persistence type of screen phosphor, so that a visual impression of the transient trace will remain continuously on the screen between cycles of the recurrent transient phenomena.

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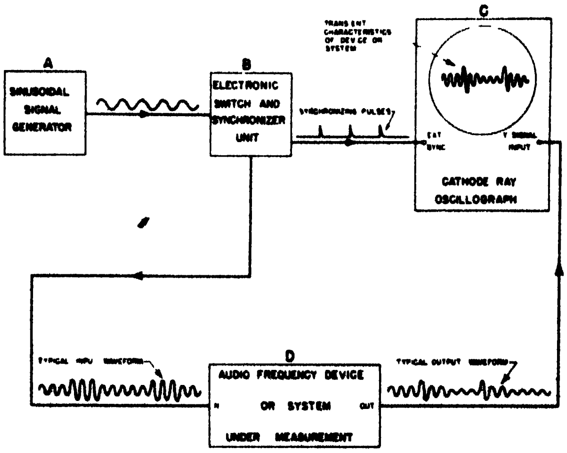
[4] Maxwell, Donald E., "Dynamic Performance of Peak-Limiting Amplifiers," *Proc. I.R.E.*, vol. 35, pp. 1349-1357, November 1947.

The phenomena which occur for the first several cycles immediately after the establishment of amplitude *b* are usually of the greatest interest. By the use of the above-described measurement technique, the gain-reducing action of the amplifier can be observed cycle by cycle of the applied sine-wave signal, and the attack time is measured by counting the number of cycles of a given frequency to the point where no further change of amplitude takes place. Accompanying undesirable effects, such as thump and waveform distortion, are also shown in as great detail as desired merely by changing the sweep speed of the oscilloscope.

An important requirement of this method of transient analysis is that the change from amplitude *a* to amplitude *b* of Fig. 15-105 be made at a time when the applied sine-wave signal is crossing the axis of zero amplitude. If the amplitude change were to be made at any other part of the cycle, an irregular wavefront would be developed, rendering the transient analysis more difficult of interpretation.

Measuring Equipment

A block diagram of the required measuring equipment is shown in Fig. 15-106. A signal generator *A*, which may be a standard commercial type of audio oscillator, delivers a sinusoidal voltage of predetermined frequency to an electronic switch and synchronizer unit *B*. The electronic switch and synchronizer unit consists of a combination of electronic circuits which provide the signal-amplitude changes, phasing, and synchronizing functions required for the visual presentation of the transient phenomena on the cathode-ray oscilloscope *C*. The electronic switch and synchronizer unit supplies the input terminals of the peak-limiting amplifier under measurement (block *D* of Fig. 15-106), with voltage of the waveforms shown in Fig. 15-105. It also supplies a synchronized triggering voltage to the linear-sweep-control circuits of the cathode-ray oscilloscope *C*. The oscilloscope may be a standard commercial unit, providing it is designed for a single-sweep, externally triggered operation. The transient waveforms displayed on the screen by this method lend themselves readily to photographing. The photographs, which form the basis of discussion, were selected from a considerable number taken of peak-limiting-amplifier performance under various operating conditions.

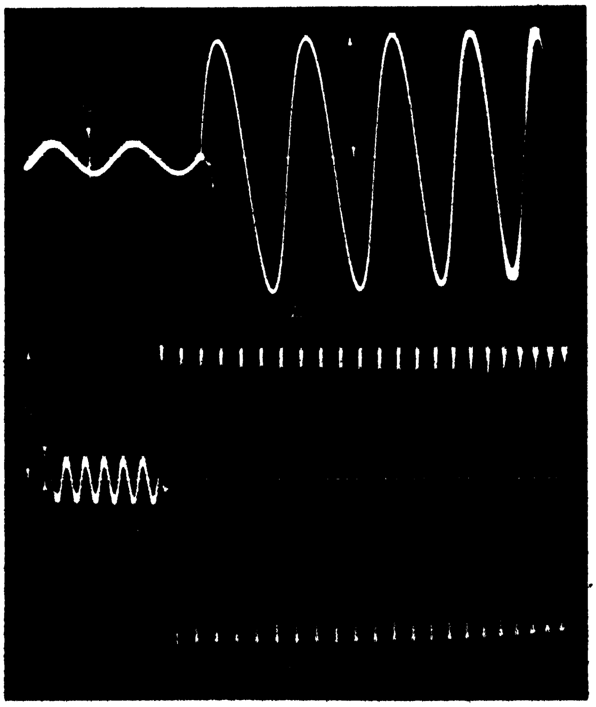

A
SINUSOIDAL SIGNAL GENERATOR
B
ELECTRONIC SWITCH AND SYNCHRONIZER UNIT
SYNCHRONIZING PULSES
TRANSIENT CHARACTERISTICS OF DEVICE OR SYSTEM
C
EXT SYNC
Y SIGNAL INPUT
CATHODE RAY OSCILLOGRAPH
TYPICAL INPUT WAVEFORM
D
AUDIO FREQUENCY DEVICE OR SYSTEM UNDER MEASUREMENT
IN
OUT
TYPICAL OUTPUT WAVEFORM

Courtesy Proc. I.R.E. and CBS

Fig. 15-106.—Block diagram of equipment setup for dynamic measurements on peak-limiting amplifiers used in broadcasting.

Observed Dynamic Performance of Commercial Peak-Limiting Amplifiers

The equipment described in the preceding section permits analysis of peak-limiting-amplifier performance at any frequency between 100 and 15,000 cps, and at any desired ratio of peak signal level to threshold signal level (ratio of amplitude *b* of Fig. 15-105 to amplitude *a*). For the sake of brevity, however, results shown in this discussion have been confined to two frequencies, 1,000 and 10,000 cps, and a single peak-to-threshold signal-amplitude ratio of approximately 18 db. The photographs, shown in Fig. 15-107, are the



Courtesy Proc. I.R.E. and CBS

Fig. 15-107.—Transient signal waveforms: (A) 1,000-cps input; (B) 10,000-cps input.

transient signal waveforms which apply successively to the input terminals of each of five different peak-limiting amplifiers. The application of these waveforms to the input terminals result in the output waveforms analyzed individually below for each of the amplifiers. Fig. 15-107A shows the 1,000-cps input waveform, while (B) shows the 10,000-cps waveform. These photographs are direct time exposures of the phenomena displayed on the cathode-ray-tube screen. For the purpose of analysis and discussion, reference axes and boundaries have been hand-drawn on many of the photographic prints shown in Figs. 15-107 through 15-111.

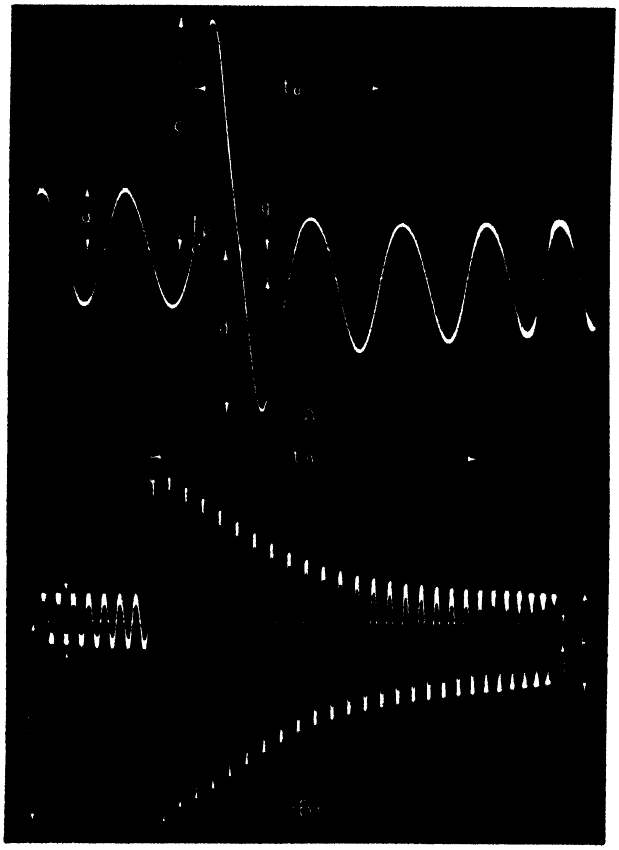
The measuring equipment is so adjusted that some 2 to 5 cps immediately preceding the arrival of the transient peak signal are shown for purposes of comparison with phenomena occurring after the arrival of the transient. As mentioned above, the high-amplitude transient signal (amplitude *b* of Fig. 15-107A and B persists for about 10 milliseconds, which is equivalent to 10 cps of the 1,000-cps signal, and 100 cps of the 10,000-cps signal. The sweep speed of the oscilloscope is adjusted for good resolution of each individual cycle, which usually results in only the first few cycles, immediately after the onset of the peak, being displayed on the screen. This is illustrated by Fig. 15-107A, where 2 cps before and 4 cps after the onset of the peak appear on the screen of the cathode-ray tube. Obviously, merely by changing the sweep speed of the oscilloscope, each cycle can be studied in as great detail as desired, or, if the sweep speed is made sufficiently slow, the action of the peak-limiting amplifier may be observed throughout the entire 10-millisecond period. In general, a relatively fast sweep speed is used when it is desired to study waveform distortion in detail, and a slower sweep speed is used when the peak envelope is of chief interest. The case of a relatively slow sweep speed is illustrated in Fig. 15-107B for the 10,000-cps signal, where the first twenty-odd cycles of high-amplitude transient appear on the cathode-ray tube. In this latter case, any cycle-to-cycle peak-amplitude variations would be clearly indicated.

Considerable care was taken in the generation of the applied waveforms of Fig. 15-107 to insure that no d-c component of voltage was included in the high-amplitude signal after the points *p*. Close examination of these waveforms will reveal a slight dissymmetry of the positive and negative peak amplitudes due to second-harmonic distortion in the high-amplitude signal. This is not a desirable condition, but one imposed by signal-handling limitations in the electronic switch and synchronizer unit. Further development of the latter unit, since these photographs were taken, eliminated this distortion, but it was not considered of sufficient magnitude to affect substantially the results of the present analysis of peak-limiting-amplifier performance.

The response of several commercial peak-limiting amplifiers to the applied waveforms of Fig. 15-107 will now be analyzed. The sole purpose of these analyses is to describe dynamic phenomena which occur in typical commercial peak-limiting amplifiers; they are intended to be neither a recommendation nor a condemnation of the subject amplifiers.

Peak-Limiting Amplifier No. 1

The first two cycles shown at the left of Fig. 15-108A represent an output level corresponding to the threshold of gain reduction, prior to the onset of the sinusoidal signal peak. Study of the waveform beyond the point *p* reveals that the first half-cycle rises to an amplitude *c* approximately 12 db above the threshold amplitude, showing that gain-reducing action has been insufficiently fast to reduce the amplitude of the first



Courtesy Proc. I.R.E. and CBS

Fig. 15-108.—(A) This pattern shows that approximately two complete cycles of the 1,000-cps peak are required for effectively complete gain reduction; (B) the effects on amplifier No. 1 of a peak signal having a frequency of 10,000 cps are indicated here.

half-cycle of the 18-db, 1,000-cps peak by more than approximately 6 db. The amplitude *d* of the second half-cycle is seen to be reduced further, but it is still some 9 db above the peak threshold level *a*. It is evident from Fig. 15-108A that approximately two complete cycles of the 1,000-cps peak are required for effectively complete gain reduction. This can be interpreted to mean that the attack time of this amplifier is 2 milliseconds (the period of 2 cps of a 1,000-cps frequency), as indicated by the time t_a. It can be seen from the figure that there has taken place an axial shift of the output voltage during the process of gain reduction, as indicated by the distance *g* between the axes of the sine wave before and after gain reduction. This is the result of a d-c component of control voltage which has not been completely balanced out, and represents the "thump" component of this amplifier for the particular peak signal applied. Note that, even though the sine-wave amplitude after gain reduction is apparently no greater than before the arrival of the peak, the axial shift due to thump results in all negative peaks shown on Fig. 15-108A exceeding the negative-peak threshold amplitude *a* by a value of approximately 4 db. Therefore, were amplitude *a* equivalent to 100 per cent modulation of an associated transmitter, substantial overmodulation would persist for a much longer period than the apparent attack time of 2 milliseconds.

Since the signal path of no commercial peak-limiting amplifier can pass d.c., the time duration of the thump component indicated by *g* is a complicated function of the low-frequency response of the amplifier signal and control-voltage circuits, as well as the feedback-loop gain of the control-voltage circuit. Frequently, the amplitude of the thump component decreases in the form of a damped oscillation. This can be considered as an envelope modulation of the signal frequency, and an oscillation frequency of the order of 5 to 10 cps is common. Therefore, the 2- or 3-millisecond observation period of Fig. 15-108A is too short to indicate any appreciable decrease in the magnitude of the thump amplitude *g*.

A consideration of the above factors makes it evident that a constant output amplitude may not be reached for an appreciable fraction of a second after the application of a sustained peak signal. The actual time required for an essentially steady-state amplitude to be attained is thus a function of the original magnitude of the thump component and the decay period of the individual amplifier. This low-frequency thump component can very readily have a more disagreeable listening and operational effect on the signal than the short-duration, high-amplitude bursts which pass through the amplifier due to an insufficiently short attack time. For instance, it is not uncommon for the modulator of an amplitude-modulated transmitter using inverse feedback to have a sharply rising, subaudible, low-frequency response, the peak of which may coincide with the thump envelope frequency; in which case the thump is aggravated, and the modulator may be completely disabled for the duration of the thump. In an extreme case, there might be developed an oscillation of sufficient amplitude to trip an overload circuit and remove the transmitter from the air.

The magnitude of the thump component in any amplifier varies with the amplitude of the peak which produces gain reaction. Some peak-limiting amplifiers have a so-called thump control which is effective in balancing the thump for any single amplitude of peak signal, but since the gain-reducing circuits seldom exhibit the same degree of balance at any other degree of gain reduction, these thump controls merely permit a compromise adjustment which has the lowest average thump content under normal program conditions.

The thump amplitude *g* of Fig. 15-108A is by no means of unusual magnitude, as peak-limiting amplifiers go. In fact, it is probably not great enough to be detected solely by a listening test of the audio output of the amplifier. Nor is it likely, either, that the high amplitude bursts which are shown to occur for the first cycle or two can be detected by a simple listening test.

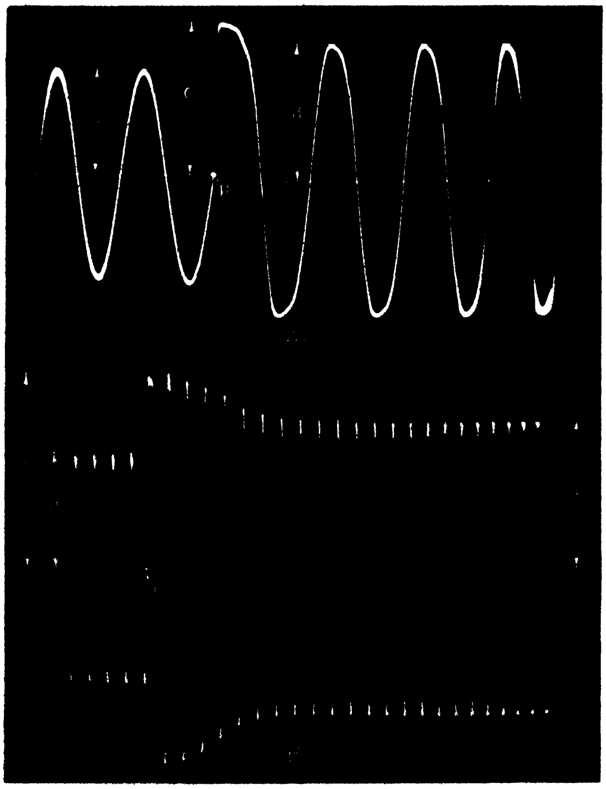
So far, no detailed tests have been made to correlate the transient effects observed on the oscillograph with the subjective listening effects of the transient phenomena. Preliminary observations, however, indicate that if the amplitude of the thump component, shown by the above sine-wave tests, is as great as the threshold signal amplitude (i.e., if amplitude *g* is as great as amplitude *a*), then a listening test with ordinary program material is very likely to disclose a disagreeable thump each time heavy gain-reducing action occurs. Perhaps even more significant than the amount of audible thump observed at the output terminals of the amplifier, however, is the possible ill effect the thump component may have on subsequent audio equipment, such as the modulator described above.

Fig. 15-108B shows the effects on amplifier No. 1 of a peak signal having a frequency of 10,000 cps. Here the 2-millisecond attack time t_a is perhaps more clearly illustrated than in the 1,000-cps case. It is seen that approximately 20 cps are required after the application of the peak for the output amplitude to approach the more or less constant value *k*. The first few cycles of the 10,000-cps peak pass through the amplifier with little or no attenuation.

It is noted that the amplitude *k* is about 6 db greater than the amplitude *a*, a very undesirable condition probably indicating deficient high-frequency response in the control-voltage circuit. This is a deficiency which would also be indicated by steady-state measurements. A thump component *g* is also present in the 10,000-cps case.

Peak-Limiting Amplifier No. 2

The photographs of Fig. 15-109 show that amplifier No. 2 has considerably less amplitude of overshoot than amplifier No. 1 during the period while gain reduction is taking place. This particular amplifier is designed to produce peak-chopping action at an output amplitude approximately 3 db above the threshold amplitude. The peak-chopping action is independent of automatic gain reduction, and, hence, limits the maximum peak amplitude to about 3 db above the threshold value, regardless of how long it takes for complete gain reduction to be effected. Amplitude *c* of Fig. 15-109A and (B) corresponds to the peak-chopping level of this amplifier, and, were there no gain-reducing action, the output waveform after point *p* would be flat-topped and of amplitude *c*. Note, in the 1,000-cps case, Fig. 15-109A, that a substantial part of the first half-cycle remains at or near the peak-chopping level, before sufficient gain reduction has occurred to reduce the output amplitude below this value. Fig. 15-109B shows that the output level remains near the peak-chopping value for about the first three cycles of a 10,000-cps peak.



Courtesy Proc. I.R.E. and CBS

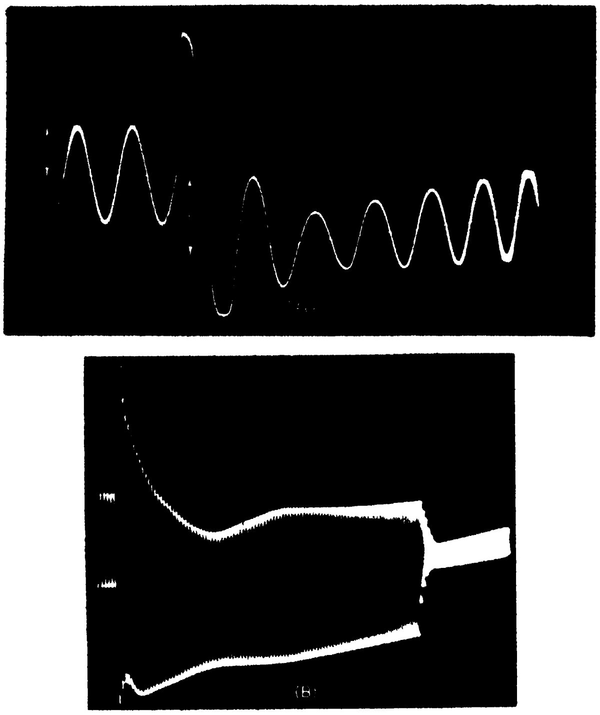
Fig. 15-109.—These photos show that amplifier No. 2 has considerably less amplitude of overshoot than amplifier No. 1 during the period while gain reduction is taking place. (A) 1,000-cps input; (B) 10,000-cps input.

It may be noted from the figure that considerable waveform distortion is evident near the peaks of at least the first four complete cycles after the arrival of the 1,000-cps peak. This indicates that gain-reducing action is still going on, even though the peak amplitude reaches a relatively stable value *d* after the first cycle. If Fig. 15-109B were expanded by a faster sweep, considerable waveform distortion would also be observable over the first 30 or 40 cps after the arrival of the 10,000-cps peak. From this figure it is seen that approximately 6 cps are required for the output amplitude to reach its stable value *d*, which indicates that the attack time of amplifier No. 2 is approximately 0.6 millisecond. In view of the fact that considerable waveform distortion persists for a much longer period, however, this factor should probably be taken into account. If the attack time of amplifier No. 2 is based upon the time required for the output waveform to become essentially sinusoidal, the value would be on the order of 4 milliseconds, instead of 0.6 millisecond.

Amplifier No. 2 appears to have a low thump component, as indicated by very little axial shift of the waveform before and after gain reduction. This particular amplifier is provided with a thump control, and optimum adjustment of this control was made before the photographs were taken. However, as noted in the discussion of amplifier No. 1, the thump control insures a low thump component for only one particular amplitude of peak signal; for some lower or higher amplitude of peak signal, amplifier No. 2 might exhibit appreciable thumps.

Peak-Limiting Amplifier No. 3

As evidenced in Fig. 15-110, this amplifier exhibits a heavy thump component. The thump effect of amplifier No. 3 is great enough to be definitely audible on a listening test with ordinary program material. It is observable from part (A) of the figure that the initial thump (amplitude *c*) is almost twice the peak threshold amplitude *a*. Note that the zero-signal axis, after the application of the 1,000-cps peak, has a definite upward slope to it. This indicates that amplifier No. 3 has a more rapid thump-decay period than was exhibited by amplifier No. 1.



Courtesy Proc. I.R.E. and CBS

Fig. 15-110.—This figure indicates a heavy thump component for peak-limiting amplifier No. 3. (A) The initial thump amplitude *c* is almost twice the peak threshold amplitude *a*; (B) 10,000-cps output when the waveform of Fig. 15-107B is applied to the input.

Fig. 15-110A illustrates over-control, or excessive gain reduction. It can be observed that the fourth half-cycle has an amplitude (as also have the next several cycles thereafter) *less* than the threshold amplitude *a*. This is a direct result of too heavy a thump component, and the thump voltage is rectified along with the signal voltage, producing excessive control voltage in the gain-reducing circuit.

At the end of the 10-millisecond peak, the output voltage of the amplifier is seen to be less than that immediately preceding the onset of the peak, the difference between the two amplitudes being a measure of the gain reduction that has taken place in the amplifier.

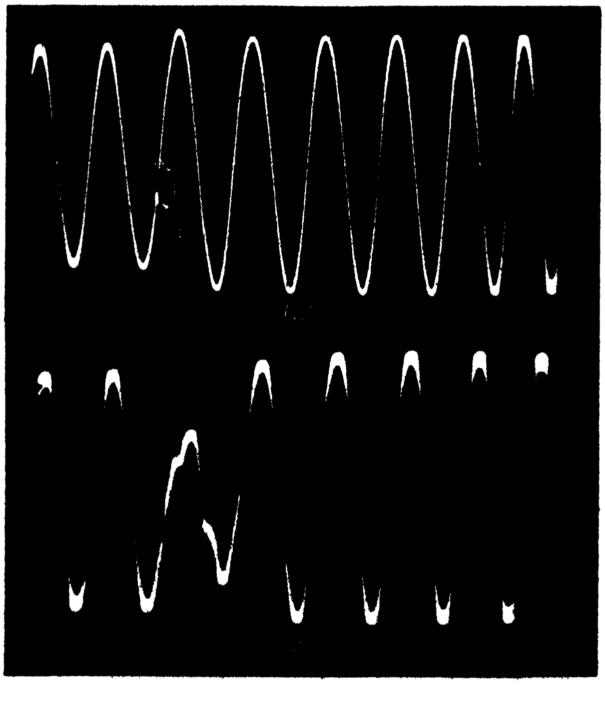
Due to the severe thump components of amplifier No. 3, it is difficult to specify its attack time definitely. For instance, although Fig. 15-110A indicates that maximum gain reduction occurs about 2 milliseconds after the application of the peak, the negative peak amplitude at that point is still more than 6 db greater than the threshold amplitude *a*, due to the magnitude of the thump. Nor has a stable output amplitude been reached within the 10-millisecond period of duration of the peak.

Fig. 15-110B shows the 10,000-cps output when the waveform of Fig. 15-107B is applied to the input. This photograph differs slightly from the 10,000-cps photographs of Figs. 15-107, 15-108, and 15-109 in that a slower sweep speed has been employed. In this photograph, the entire 10-millisecond period of duration of the peak is visible. So close are the individual cycles under this condition that the resulting picture is essentially an envelope of the output peak amplitudes. It is difficult to offer a rational explanation for the shape of this 10,000-cps envelope, since it is such a complicated function of the transient characteristic of the amplifier circuits.

CBS Experimental Amplifier

In one or several ways, it has been observed that each of the peak-limiting amplifiers analyzed leaves considerable room for improvement. The Columbia Broadcasting System recently developed a unit which has outstanding dynamic-performance characteristics.

The CBS amplifier has a maximum control range of 14 db; therefore, the analyses of Fig. 15-111 differ from those covered in Figs. 15-107 through 15-110 in that the transient peak amplitude is 14 db above the threshold value rather than 18 db. The 14-db transient was simulated at each of two frequencies, 1,000 and



Courtesy Proc. I.R.E. and CBS

Fig. 15-111.—Output waveforms of CBS amplifier: (A) 1,000-cps dynamic performance; the first two cycles at the left show the output voltage immediately before the arrival of the transient peak and the succeeding six cycles correspond to the output voltage immediately after the arrival of the peak; (B) 10,000-cps performance illustrates the extremely short effective attack time of the amplifier.

10,000 cps. Except for the 4-db difference in amplitude, the input waveforms were similar to those shown in Fig. 15-107 and the resulting output waveforms of the CBS amplifier are shown in Fig. 15-111.

With reference to the 1,000-cps dynamic performance shown in (A) of the figure, the first two cycles at the left of the photograph show the output voltage immediately before the arrival of the transient peak, while the next six cycles correspond to the output voltage immediately after the arrival of the peak. Note that not even the first half-cycle after the arrival of the peak shows any appreciable amplitude overshoot. Thus the attack time of the CBS amplifier under the above conditions is effectively zero. What may seem more surprising is the fact that there is little waveform distortion of even the first quarter-cycle after the arrival of the 1,000-cps transient.

This excellent performance is attributable to a unique circuit design wherein the automatic-gain-control voltage is a function of the input signal voltage, rather than of the output signal voltage. The control-voltage generating section of the amplifier incorporates large power-type amplifier and rectifier tubes in low-impedance circuit arrangements, resulting in extremely fast development of the automatic control voltage. Another major factor contributing to the exceptional performance shown in the figure is the use of a special time-delay network in the signal channel just ahead of the point where gain reduction takes place. This network acts to delay the signal by approximately 80 microseconds, and the gain is already reduced by the required amount upon the arrival of the peak at the point where gain control is effected. A close examination of Fig. 15-111A will reveal that automatic gain reduction occurs near the point *q* in time, whereas the 14-db peak does not arrive until a later time at point *p*.

The 10,000-cps performance shown by Fig. 15-111B again illustrates the extremely short effective attack time of the amplifier. Even at 10,000 cps, the 14-db peak which arrives at the point *p* in time is effectively prevented from exceeding the maximum steady-state value. The figure indicates excessive gain reduction (over-control) for the first cycle after the arrival of the peak. Over-control which exists for so short a time as that shown (approximately 100 microseconds, or one cycle of a 10,000-cps wave) cannot be perceived by a listening test, and is certainly preferable to under-control since it renders overmodulation of subsequent equipment impossible.

It can be observed from Fig. 15-111 that the "thump" component of the amplifier is of very small amplitude, an additional design feature of this unit.

The amplifier described above was developed by E. E. Schroeder of the CBS-Chicago technical staff, under the direction of J. J. Beloungy, and has been used at station WBBM since 1945. Additional amplifiers, based upon this development, are in service in other Columbia Broadcasting System stations and are also available commercially from a well-known manufacturer.

CHAPTER 16

VISUAL ALIGNMENT OF A-M, F-M, AND TELEVISION RECEIVERS

Every receiver requires that its tuned circuits be correctly adjusted for optimum operation. Since it is always possible for a tuned circuit to be thrown off its proper resonance by a number of factors, either an inductive or a capacitive component is made variable, so that the proper resonant point can be maintained. The process by which the circuits are adjusted to the proper resonant state is called *alignment.*

Alignment is one of the most important factors in the servicing of receivers, because the receiver will not function properly if it is incorrectly aligned. Knowledge of alignment of receivers dates back to the time of tuned r-f sets. However, as the years advanced, especially with the coming of the superheterodyne receiver, the alignment procedure became more involved and also more necessary. Today we actually have three different types of superheterodyne receivers to service, namely, a-m, f-m, and television. Because of these last two later types of receivers, alignment processes are not as simple as they used to be and are considered even more important than before. When the radio market consisted primarily of a-m receivers, alignment procedures usually involved the use of a signal generator and some form of output indicating device, such as a simple output meter or VTVM. Correct alignment was usually indicated by a maximum reading on the meter. Besides this meter method of alignment, the oscilloscope, in conjunction with a special type of signal generator, was also used as the output device for *visually* aligning the receiver. At the beginning, this visual alignment method was not very popular with the serviceman primarily because of the expense involved in buying the necessary equipment. Today, because of the increasing production of f-m and television receivers, visual alignment is considered an essential part of servicing. The design of equipment needed for visual alignment has been greatly advanced and the cost of instruments very much reduced so that they are within the reach of most servicemen. In most instances, the visual alignment method is considered far superior to the meter method.

Just as a casual comment, let it be known that visual curve-tracing devices, such as the oscilloscope, when used for alignment operation, have been in use for some time. While it is true that all of these commercial visual curve-tracing devices were not of the cathode-ray-tube variety, they were oscillographic. Incidentally, as far back as 1912, the oscilloscope was used with a motor-driven capacitor and a synchronized potentiometer-battery type of sweep for the tracing of resonance curves. Deflection coils were used in the sweep circuit instead of deflection plates. In the early 1930's, the production lines of many receiver manufacturers used the string galvanometer type of oscillograph, modernized of course, for the alignment and testing of the various tuned circuits which were part of the receivers being produced.

Basic Facts About Visual Alignment

The alignment of a tuned circuit, the establishing of the correct resonant state, by visual observation is commonly understood to be one of the most important servicing applications of the oscilloscope. Having read through this volume up to this point, you no doubt appreciate our statement when we say that observation of the response of resonant circuits is but one more application of the cathode-ray instrument we are discussing. It is true, however, that from the viewpoint of the servicing and manufacturing industries, it is an extremely important application. In fact, many service technicians have displayed infinitely more interest in its uses for the purpose of tuned circuit alignment than in its myriad other applications.

If we analyze the basic principles of the curve-tracing or visual-alignment processes employed in connection with cathode-ray tubes, we find them very much like those which apply to other applications of the oscilloscope. It is true that there are various ways of accomplishing some of the actions which enable operation of the complete visual-aligning system, but the basis of the system has already been described, although it may not have appeared as such.

If we remember that the image which appears upon the screen of the cathode-ray tube, irrespective of its shape, is the result of a deflection of the beam by means of a varying voltage applied to the deflection plates, we have the basis for a visual-alignment system. This, as you know, is also the basis for the normal operation of the oscilloscope.

There are, however, two additional considerations found in alignment systems which are not involved in the simple waveform observation operations with the oscilloscope. One of these is that instead of a fixed frequency voltage input to the device under test, as is the case when observing voltage waveforms, the frequency of the voltage input *varies continuously over a predetermined band*. This frequency varying voltage, or *frequency-modulated (f-m) signal*, as it is often called, is necessary in order to visualize the resonant curve inasmuch as such curves cover certain bandwidths. This f-m wave is obtained today from special alignment generators called *sweep generators*. (See Chapter 14.) After passing through the network under test, the signal is rectified and then applied to the vertical-deflection terminals of the oscilloscope.

The second consideration is that relating to the time-base signal that is to be applied to the horizontal-deflection plates of the oscilloscope. We do know that when the oscilloscope is used for normal waveform observation, the waveshape of its time-base signal is linear (a sawtooth wave) and that the frequency of this signal is adjusted to a certain ratio with the frequency of the voltage being observed, so as to place a certain number of cycles upon the screen. For visual-alignment work, the situation is completely different, the shape of the time-base signal used by the oscilloscope does not have to be linear but must bear a definite relationship to the modulating signal producing the f-m wave used in this process of alignment.

Besides developing the response curve of the tuned circuit on the oscilloscope, the process of visual alignment has advanced to the point where it is a simple matter to spot frequencies along the curve. The equipment used to produce this latter effect is commonly known as a *marker* generator.

In Fig. 16-1 appear three block diagrams illustrating the basic setup for visual alignment. Part (A) shows the sweep generator and marker generator as parts of a single piece of equipment and their combined signals fed to the receiver under test. The proper signal output is then fed to the vertical terminals of the oscilloscope. The horizontal-deflection voltage (i.e. the time-base signal) is normally obtained from the sweep generator. This signal is the same as that which causes the frequency modulation inside the sweep generator (i.e. the modulating signal). The second setup in part (B) merely illustrates the condition where the marker generator is a separate unit from the sweep generator. The remainder of the setup is the same as in part (A).

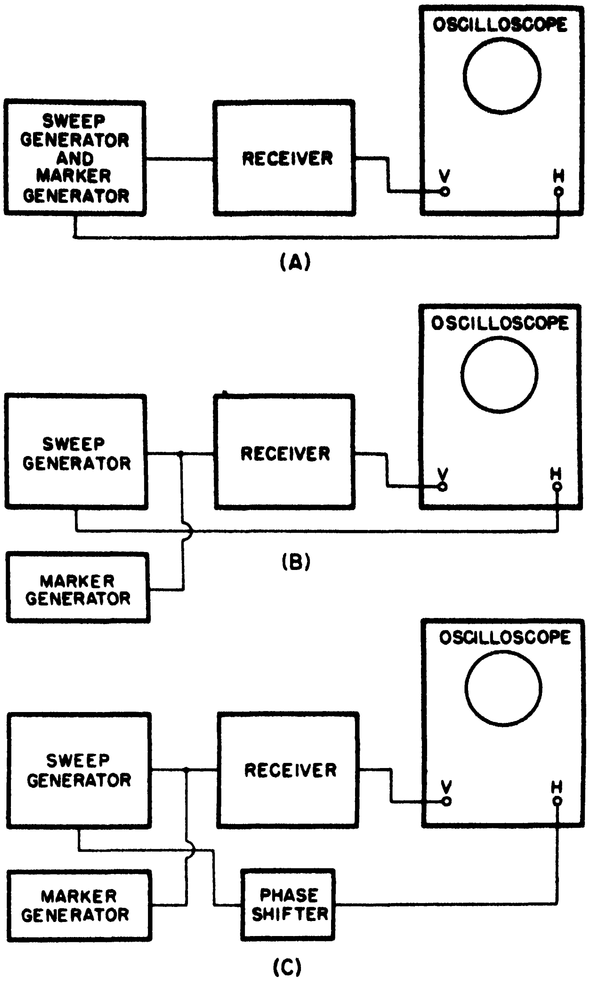


Fig. 16-1.—Block diagrams illustrating the basic equipment and hookup methods for visual alignment.

Most modern sweep generators today come equipped with a phasing control which, by proper manipulation, makes the two deflection voltages of the oscilloscope in phase with each other so that only a single trace of the pattern appears on the screen. For those sweep generators that do not contain such a control, a simple phase shifter can be inserted between the horizontal terminals of the oscilloscope and the modulating signal output terminal of the oscilloscope, as in (C).

Before we discuss the actual visual-alignment processes of the various receivers, it is best that we understand a few things about the tuned circuits of the receiver, the equipment used, the manner in which the pattern is formed, and the techniques involved in visual alignment. To start, we will first study some of the fundamental details about resonance curves and of the bandwidth and shape of the tuned circuits found in receivers.

Resonance Curves

When dealing with the visual alignment of tuned circuits, the picture seen on the screen of the oscilloscope is the response of a single tuned circuit or some combination of tuned circuits. This picture is usually termed a *resonance curve* or *response curve.* In brief, a resonance curve is a graph of the relative amplitude (either voltage or current) of the signal output from a tuned circuit or tuned system with respect to a number of frequencies on either side of the resonant frequency. When viewed on the oscilloscope, the resonance curve seen is actually a voltage versus frequency characteristic because the oscilloscope is a voltage indicating device.

Resonance curves of tuned circuits can be graphically plotted. All that is required is a signal generator (unmodulated) capable of the necessary frequency range of the circuit, and a voltmeter. The basic setup is illustrated in part (A) of Fig. 16-2 where the box indicates the tuned circuit or system whose response is to be plotted. The signal generator is placed across the input to the tuned circuit and the voltmeter placed across the output. The frequency of the signal generator is varied step-by-step over the pass-band of the tuned circuit being tested and the output voltage from the circuit measured for each frequency setting of the generator. During this complete operation the input voltage to the tuned circuits must be kept *constant* during all frequency settings. By using known values of the frequency input as the abscissa or X-axis and the indicated values of the voltage output as the ordinate or Y-axis these points can be plotted and then all joined together by a single continuous line. Part (B) of Fig. 16-2 illustrates such a plot where the resulting curve is the response of the tuned circuit.

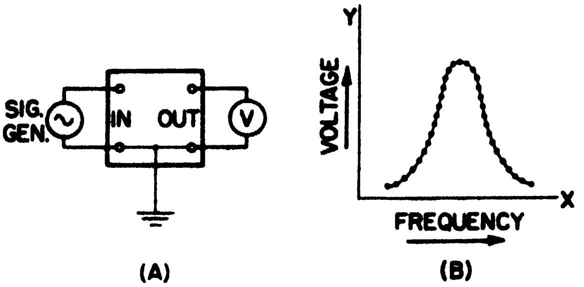


Fig. 16-2.—(A) Basic setup for point-by-point checking of the response of a tuned circuit, (B) response curve obtained by this method.

Such a resonance curve as shown in Fig. 16-2B can represent that of a single- or double-tuned parallel L-C circuit. For such a circuit, the resonant frequency results when the output voltage is a maximum and occurs at the peak of the resonance curve. The curve is usually symmetrical about the line of resonant frequency. To obtain a resonance curve by this point-by-point method is quite tedious and time consuming and, therefore, impractical to use. By the use of a frequency-modulated signal generator (i.e. sweep generator) and an oscilloscope, the actual resonance curve can be seen without recourse to plotting. Very little time is taken, for this visual procedure as compared to that taken for the point-to-point graphical plot system.

If a tuned circuit or stage is to be aligned then its variable circuit component is normally adjusted for maximum response, *or as indicated by the manufacturer's specific instructions.* In many types of equipment, it is possible to use a meter method to align certain stages. In general, if the output meter indicates a maximum then the circuit is assumed to be tuned to resonance, that is, it is aligned. This method is all right as far as simple, single-peaked tuned circuits are concerned and has been used quite extensively in the past for a-m receiver alignment. However, with f-m and television receivers the tuned circuits have become more complicated to align and use of the meter method has its limitations. For instance, if the tuned stage in question is an overcoupled transformer, then the response of this circuit will be a resonance curve that is double-peaked. The actual resonant point of the circuit lies in the trough between the two peaks and a maximum meter reading in this case would only indicate one of the peaks of the curve and not the true resonant point. Of course, the true resonant point can be found by this meter method; it is the minimum meter reading that lies in between the two maximum meter readings. This method is time consuming, however, and not too accurate, especially if there is very little difference between the voltage amplitude of the trough and the peaks.

In the visual-alignment process, the curve appears on the screen of the oscilloscope and the tuned circuit is adjusted to give the desired curve shape. Since alignment today generally encompasses more than one tuned circuit or stage, the curve seen on the oscilloscope usually represents a combination of individual resonance curves. Because of this latter fact, the overall resonance curve can be either symmetrical or asymmetrical. By a symmetrical resonance curve is meant one which shows the same amplitude response for frequencies which are higher and lower by equal amounts

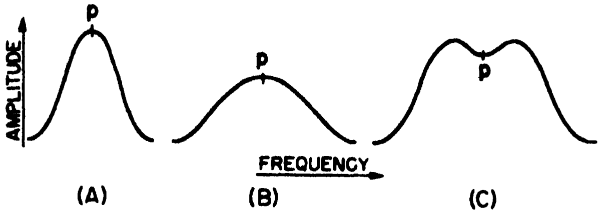


Fig. 16-3.—Three typical symmetrical resonance curves.

than the resonant frequency. Another definition of a symmetrical resonance curve is one in which each half of the curve about its resonance line is a mirror image of the other half. Asymmetrical curves, also often called nonsymmetrical curves, do not possess these characteristics. A slight deviation in the shape of a symmetrical curve may make the curve asymmetrical.

Resonance curves may be of various shapes, yet be symmetrical. In other words, symmetrical resonance curves may be obtained with loosely coupled circuits, critically coupled circuits, and overcoupled circuits. Resonance curves may be single-peaked, double-peaked, or even triple-peaked in shape. Fig. 16-3 illustrates three typical types of symmetrical resonance curves encountered in alignment work. Each one of these three curves can be obtained by a single-tuned stage. That at (A) is a single-peaked sharply resonant curve which may result from a high *Q* circuit, and that at (B) is also single-peaked but broadly resonant, which may result from a low *Q* circuit. Both types of curves are obtained from loosely coupled circuits. From the shape of these two curves, it is readily seen that curve (A) has a higher selectivity than curve (B). The response at (C) is a double-peaked curve and is usually produced by overcoupled circuits. Note how these three curves vary in shape but that each still possesses symmetry about its resonant frequency, indicated by point *P* in each drawing. Such curves can also be the over-all response of a number of tuned circuits. If each tuned circuit in this latter instance has the same resonant frequency then *P* can still be said to represent the resonant frequency of the curves.

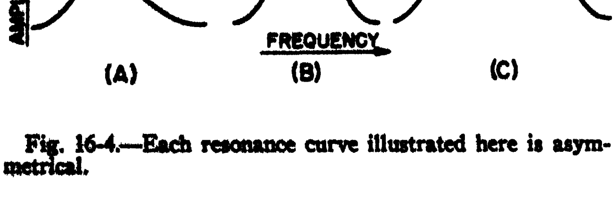


Fig. 16-4.—Each resonance curve illustrated here is asymmetrical.

Three different asymmetrical resonance curves are shown in Fig. 16-4. Part (A) represents a single-peaked asymmetrical curve while parts (B) and (C) are double-peaked curves. These curves can also be the response of either a single-tuned circuit or a combination of tuned circuits. If they are the result of a single-tuned circuit then the double-peaked curves must result when the circuit is overcoupled. However, if the responses of curves (B) and (C) are the results of a combination of tuned circuits, the individual resonance curves of each tuned circuit may have a single peak (i.e. be loosely coupled). Their respective shapes and resonant frequency points may be such that their combination results in the curves of parts (B) and (C).

In certain cases, it is merely desired to broaden the response of a single-tuned circuit or system; the exact shape is secondary as long as the curve will accept frequencies within certain limits. In other cases, the shaping of the response curve is extremely important. This is so because proper operation of the receiver is based upon this shaping. This is especially so in f-m and television receivers. The two response curves of Fig. 16-5 illustrate two of the most important cases of shaping. Part (A) of this drawing is the required response curve of the detector system of an f-m receiver or the f-m sound section of a television receiver. Besides having the proper bandwidth between points *a* and *b* on this curve the shape of the curve between these two points must be *linear* for proper detection of f-m signals.

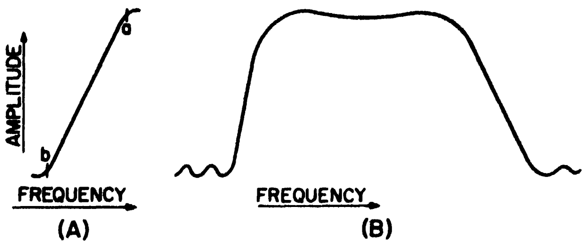


Fig. 16-5.—The typical f-m detector curve (A) and the video i-f curve of (B) indicate two important cases where the curves have to be of prescribed shapes.

Part (B) of Fig. 16-5 is a typical response curve of the over-all video i-f system of a television receiver. The individual video i-f stage may be overcoupled or loosely coupled thereby having either double-peaked or single-peaked response curve. If they are single-peaked, the resonant frequency of each i-f stage is usually not the same, that is, they must be stagger-tuned. This is necessary in order to obtain the proper total bandwidth. Besides the video i-f transformers, in those television receivers that employ video and i-f wave traps, these traps also have to be adjusted to help produce the desired pattern. In the shaping of the curve of Fig. 16-5B both the right and left sides of the curve warrant attention. The steepness of the slopes should be approximately as indicated in the drawing. Proper adjustment of the traps helps produce the desired slopes and also the shaping of the ends.

Without previous knowledge of the resonant frequencies or the shape of the response curves of tuned circuits that are to be aligned, it is difficult to tell exactly what to look for. We do not know if the required resonance curves are to be symmetrical or asymmetrical. For the alignment of simple symmetrical single-peaked tuned circuits, the resonant frequency is usually all that is needed; the knowledge that the circuit will produce a symmetrical single-peak resonance curve is sufficient. The same is true for symmetrical double-peaked curves. However, where the desired curve is asymmetrical, whether it be the response of an individual transformer-coupled circuit or a combination of different tuned circuits, the desired shape of the curve must be known.

Whatever the case may be, it is always best to know the shape of the resonance curve and resonant frequencies of the tuned circuits in order to facilitate visual alignment. Today most f-m and television receiver manufacturers indicate, in the service notes accompanying their receivers, the shape of the resonance curves desired for proper visual alignment as well as the resonant frequencies of their tuned circuits. It is, therefore, always advisable to follow the manufacturer's notations when aligning.

So much for the general details concerning the shape of the resonance curves and their connection to the different receivers. Let us now study something about the bandwidths of tuned circuits.

Bandwidths of Tuned Circuits

Tuned circuits are frequency selective circuits in that they will pass certain frequencies and reject others. The various shapes of the resonant curves illustrated in Figs. 16-3, 16-4, and 16-5 readily indicate this frequency selectivity. All the tuned circuits that are considered in the alignment of receivers are *band-pass* circuits. By this we mean they accept a certain band of frequencies and reject signals on *either* side of this band. This so-called acceptable band of frequencies is generally referred to as the *bandwidth* of the tuned circuit. The bandwidth referred to here may encompass the frequency limits of a single-tuned circuit or a combination of tuned circuits. The bandwidth of a circuit or network is, therefore, a means of expressing the difference between its limiting frequencies. Thus if a tuned circuit or system is said to have a bandwidth of 50 kc and the actual center frequency of this circuit's response is equal to 900 kc then the limiting frequencies of this circuit are said to be 925 kc (900 + 25) and 875 kc (900 − 25).

The determination of what the bandwidth of a tuned circuit should ideally be is based upon the type of signal that is to be passed. In this respect it is the sidebands of the signal which are the most important controlling factors of the bandwidth. The sideband frequencies in a-m, f-m, and television broadcasting are quite varied so that for each type of receiver different bandwidths exist. This will be discussed in a few moments but before we go any further it would be important to mention that the definition of bandwidth as used today varies. It is generally understood to mean the frequency difference between the two limiting extremes of the circuit or curve. The important question is what determines these so-called limiting frequencies?

According to common accepted engineering practice, these limiting points are determined directly from the resonance curve of the circuit or network in question and normally are chosen as those points on either side of the curve where the voltage has dropped to 70.7 per cent of its maximum value. These points are also known as the 3-db or half-power points. Today, however, when referring to the bandwidth of a tuned circuit or system this definition of the term is not always adhered to. That is, when bandwidth is sometimes referred to, the frequency limits are not necessarily determined by the half-power points. Today it is a matter of design of the tuned circuits as to what is considered the acceptable range of frequencies the circuit will pass.

Let us refer to Fig. 16-6 for a moment. The picture shown here illustrates a typical single-peaked resonance curve. Points A and B on this curve are the half-power or 3-db markings. Frequencies f_a and f_b are the limiting frequencies as determined by the half-power points on the resonance curve. The bandwidth of this curve, according to the engineering definition, is the difference between the limiting frequencies f_a and f_b. This does not mean that frequencies on either side of f_a and f_b will not be passed. In fact frequencies between points f_x and f_y will also be passed by the tuned circuit in question but their relative signal strengths at the output of the tuned circuit will be less than those between frequencies f_a and f_b. With many tuned circuits today frequencies that fall above and below those determined by the half-power points are also considered

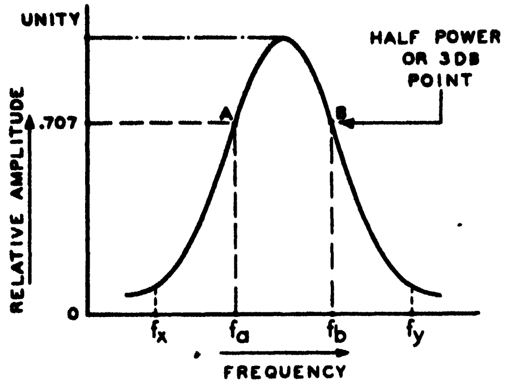


Fig. 16-6.—A typical single-peaked resonance curve indicating bandwidth markings.

"acceptable" because even though their relative signal strengths are below that at the half-power points they are strong enough to accomplish certain desired results. This applies not only to a response curve of the shape illustrated but to any shape curve. Consequently, when the term bandwidth is said, in this chapter, to be so many cycles wide it will mean that it is determined by what is considered the acceptable range of frequencies for the particular design.

The bandwidths of tuned circuits in a-m, f-m, and television receivers differ considerably from each other. In a-m broadcasting, it is the *frequency* of the audio-modulating signal which determines the minimum required bandwidth of the tuned circuits that should be used in a-m receivers. According to the FCC, 7,500 cps is the normal *maximum* audio-modulating frequency that can be used. Therefore, the normal maximum bandwidth in the a-m broadcast band is considered to be 15 kc.[1] There are not many a-m stations that modulate with frequencies as high as 7.5 kc but for comparison purposes we will consider 15 kc as the response for a-m broadcasting.

In order to make sure that the r-f section of a-m receivers accepts whatever signal the set is tuned to, the bandwidth of the r-f tuned circuit is usually made larger than 15 kc. This is necessary because in tuning from one station to another (the variable component can be either inductive or capacitive) the Q of the r-f circuit will change and hence the shape of the resonance curve will also change. Thus if the r-f circuit bandwidth is made broad enough any selected station signal will be accepted by the tuned circuit even though the tuning may change the Q of the circuit. The minimum bandwidth for the i-f circuits should, therefore, be approximately 15 kc although in practice high-fidelity receivers are the type that may have such bandwidths.

In f-m broadcasting the situation is much different. The frequency of the audio-modulating signal determines only the rate of change of the frequency swing but it is the *amplitude* of the modulating signal that determines the amount of frequency swing of the f-m signal. The FCC has standardized 100 per cent modulation in f-m broadcasting as 75-kc peak frequency swing of the carrier. Thus the maximum strength of the audio-modulating signal is limited to that producing a peak frequency swing of 75 kc.

The separation between channels in the f-m broadcast band is 200 kc. This is obtained by the FCC ruling that, to this maximum peak frequency swing of 75 kc, a 25-kc guard band is added to insure against interaction between any two adjacent stations. Thus we see that the carrier frequency of each station is separated by 100 kc on either side for a total of 200 kc. The r-f tuned circuits of f-m receivers are usually made much broader than 200 kc (400- to 600-kc bandwidths are not too uncommon) for the same reasons as mentioned for a-m receivers. The minimum required bandwidths for the i-f and detector circuits are normally 150 kc for acceptance of the 75-kc maximum peak frequency swing. In order to insure against oscillator detuning in the receiver and to make sure all the effective sidebands of the f-m signal are accepted, the bandwidths of the i.f. and detector tuned circuit are normally made between 200 and 300 kc wide. For example, the bandwidth referred to for f-m detector circuits exists between points *a* and *b* of Fig. 16-5A.

In television the bandwidths that we have to deal with are much greater than those of a-m and f-m broadcasting. The minimum frequency separation between each television channel is 6 Mc. This 6-Mc separation takes into account both the sound f-m signal as well as the video a-m signal. The r-f or input circuits of television receivers have to accept both the video- and sound-modulated signals and, therefore, the required minimum bandwidth for the input circuits is 6 Mc. The f-m signal itself occupies but a small portion of the over-all 6-Mc television channel (see Fig. 14-24 in Chapter 14). The r-f tuned circuits are usually made much broader than the minimum 6 Mc for the same reasons as given in the a-m receiver case.

The i-f tuned circuits of the television receiver are divided into two sections — the video and sound systems. It is the video i-f system which requires more careful alignment than the sound section. The shape of the over-all video i-f response characteristic is very important and, in general, should be approximately the same for all receivers. It does not matter what type of coupled circuits are employed for each individual i-f circuit as long as their combined resonance curve produces the required shape. The bandwidth of the combined video i-f circuits is also 6 Mc. A typical example appeared in Fig. 16-5B. The bandwidths of the individual video i-f tuned circuits are not the really important factor; the shape and bandwidth of the over-all curve is the important criterion. However, it is important to know the shape of the curve that appears on the oscilloscope as we progressively align *each* stage to produce the final resultant curve.

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[1] The FCC states that modulation at audio frequencies above 7,500 cps is permissible. However, if a station that modulates above 7,500 cps interferes with reception of signals from other stations, the interfering station has to install equipment or make adjustments to reduce this interference so that it is no longer objectionable. Thus in a-m broadcasting, 7.5 kc is considered as the normal maximum audio-modulating frequency.

The FCC standardizes 25-kc deviation of the sound f-m signal for television as representing 100 per cent modulation. This means that the minimum bandwidth required by the i-f and detector circuits of the sound system of television receivers is only 50 kc. The actual bandwidths of these tuned circuits are found to vary anywhere from 50 to 200 kc and possibly even higher. This increase in bandwidth is necessary in order to compensate for possible drift in the local oscillator frequency. If the bandwidths of these sound i-f circuits are about 50 kc and if the oscillator drifts, then the resultant i-f signals may be shifted part or all of the way out of the acceptable range of the i-f circuits.

So much for the basic knowledge of the tuned circuits of receivers. Now let us discuss something about the equipment used for visual alignment.

Production of the Frequency-Modulated Signal

Since the sweep generator is considered such an essential part of visual alignment it is felt that an analysis of how the frequency-modulated signal (i.e. f-m signal) is produced by such units warrants discussion. In this section, we will study those basic methods as employed in the sweep generators.

Since a resonance curve is an indication of the manner in which a tuned circuit or system responds to an impressed voltage over a band of frequencies, for visual alignment it is necessary that the input signal voltage be available as a variable frequency voltage to produce this curve. This variable frequency signal should be such that its frequency variations more than cover that represented by the extremes or bandwidth of the response curve of the circuit or system under test. If the circuit under test is that of a single resonant network or of a double-tuned transformer network, then it is said that the variable frequency signal should be changing in frequency by amounts approximately equal on both sides of the resonant frequency of the circuit. In other words, the center frequency of the variable frequency signal should be approximately equal to the resonant frequency of the tuned circuit. If the final response curve is a combination of individual resonance curves then the center frequency of the variable frequency signal should be approximately equal to the frequency that is the center of this combined response curve.

This variable frequency or frequency-modulated signal can be produced in a number of different ways. Today instruments that are used to produce the necessary f-m signals are, as we have previously mentioned, usually called sweep generators; years ago the instruments were termed wobbulators or just simply frequency modulators. One thing that all these different instruments had in common is that each contained a basic *L-C* oscillator circuit and the effective inductance or capacitance of the oscillator's tank circuit was varied at some specific rate to produce the f-m signal. It is the method of varying this effective inductance or capacitance which makes one instrument different from another. In order to fully understand the use of an f-m signal for alignment purposes we will consider some of the basic methods for the production of such a signal.

The r-f oscillator can be frequency-modulated by electronic means or by some electromechanical method. In the electronic systems there is one very important means, and that is through the use of a special circuit called the reactance-tube modulator. In the mechanical systems some component is subjected to a mechanical variation which, in turn, causes a change in the effective capacitance or inductance of the oscillator circuit.

Basic Frequency Modulator Circuit

In order to see how an oscillator circuit can become frequency-modulated let us study the simple circuit of Fig. 16-7. Here we see a capacitor-type microphone *M* shunted directly across the *L-C* tank circuit of a simple Hartley oscillator. Without any sound input to the microphone, the capacitance of the complete circuit depends upon the setting of *C* plus the fixed capacitance offered by the microphone. Thus the fixed or resting frequency of oscillation is determined by the capacitance offered by the unexcited microphone in conjunction with *L* and *C*. When a sound wave strikes the microphone the capacitance of this microphone changes. The variation is such that the capacitance changes by an equal amount above and below its fixed

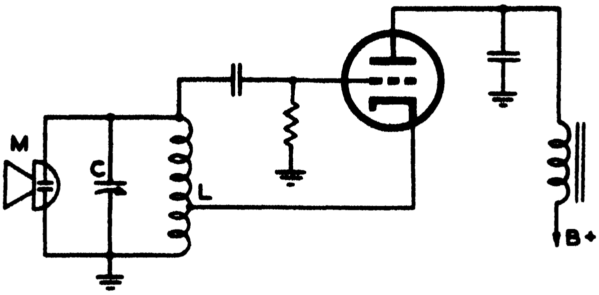

Fig. 16-7.—A simple frequency modulator circuit using a capacitor-type microphone.

value. Consequently, we can readily see that such a sound wave will cause the frequency of the oscillator to change in accordance with the changing capacitance of the microphone and hence the oscillator becomes frequency-modulated.

The sound wave is the modulating signal and the rate at which the frequency of the circuit changes about its center or resting value is determined by the frequency of the applied sound wave. The intensity or amplitude of the sound wave determines how great the change in capacitance of the microphone will be and hence the amount of frequency change of the oscillator. The stronger the sound signal the greater will be this frequency change. The amount of frequency change is limited by the physical construction of the microphone.

The amount of frequency variation about the resting frequency of the oscillator is termed the *frequency deviation* or frequency swing of the f-m signal. If an oscillator having a center frequency of say 100 kc is frequency-modulated so that the maximum instantaneous frequency is equal to 110 kc and the minimum instantaneous frequency equal to 90 kc, the total frequency deviation is equal to 20 kc. This total frequency deviation is often termed the peak-to-peak frequency deviation or just peak-to-peak deviation. The frequency difference between the resting frequency and the maximum *or* minimum instantaneous frequency of the f-m signal is often termed the peak frequency deviation or just peak deviation. The center frequency of the f-m signal is also called the carrier frequency.

The capacitor microphone circuit just discussed is not intended to convey the idea that it is the type of oscillating system actually used in sweep generators. It was shown primarily to indicate a simple method whereby an oscillator circuit can become frequency-modulated by mechanical variation of an element which is part of the oscillator tank circuit. Besides showing us a basic method of producing an f-m wave this system has also given us the fundamental relationship of the frequency and amplitude of the modulating signal to the rate and amount of deviation of the f-m signal.

Old Types of Frequency Modulators

One of the first types of frequency modulators employed for visual alignment was known as a wobbulator and was commercially available in the early 1930's. One principal type consisted of a motor-driven variable capacitor which was part of an oscillator tank circuit. A simple schematic representation of such a circuit appears in Fig. 16-8. In this drawing *L* and *C* represent the tank circuit components of the oscillator and capacitor *C1* is the tunable unit, where the rotor of *C1* is ganged to the shaft of the motor. The motor in conjunction with the variable capacitor is often called the frequency modulator or just modulator of the unit. The frequency of oscillations without the motor turned on is determined by inductance *L* and the values of *C* and *C1*.

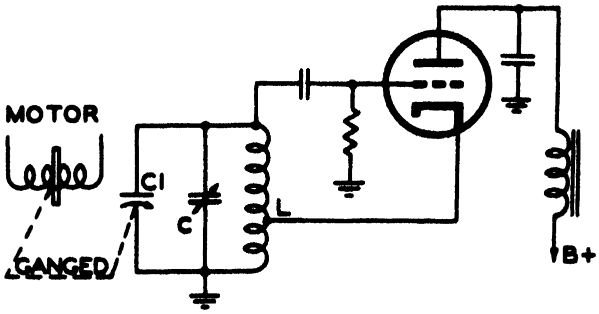

Fig. 16-8.—Schematic of one of the first types of motor-driven frequency modulators.

With the motor turned on, the rotor of capacitor *C1* is continuously rotated and since this capacitor is part of the oscillator tank circuit, each variation in capacitance causes a change in the frequency of oscillations. Therefore, as the capacitor continues to rotate, the oscillator becomes frequency modulated. The maximum and minimum capacitance values of *C1* determine the lowest and the highest instantaneous frequency of the f-m signal. The maximum amount of capacitance change of *C1* determines the frequency deviation of the f-m signal. For any one such type of wobbulator the amount of frequency deviation is fixed because the modulation capacitor was continuously rotated between its extreme capacitance values. The rate of frequency deviation depends upon the speed of the motor; consequently, the rate of frequency deviation could be varied by simply changing the speed of the motor. Primarily because of the construction of the modulation capacitors, the amount of frequency deviation produced by such wobbulators and hence the available bandwidth that the f-m signal could cover was limited. However, for visually aligning the tuned circuits of a-m receivers, where the bandwidth of the circuit is not very wide, such wobbulators found practical usefulness.

As far as the servicing industry today is concerned, motor-driven wobbulators are very seldom used, and are considered obsolete. This is especially so since the modern sweep generators are easier to use, are more versatile, and are relatively inxpensive. There are some special types, however, that are designed for certain research laboratory applications but are not commercially available.

One of the first types of electrically operated frequency modulators made use of the principle that if an alternating signal is applied across the primary of an iron-core transformer, the permeability of the iron core will change in accordance with the changing magnetizing current flowing through the primary winding. If the secondary of this transformer was connected as the inductor of an oscillator tank circuit then variation in the permeability of the iron core would cause a variation in this (secondary) inductance and hence in the resonant frequency of the oscillator circuit.

Modern Sweep Generators

Visual alignment is made extremely useful today by the modern type of frequency modulators which are better known as sweep generator.[2] Two principal types of sweep generators are used today, one where the inductance or capacitance is electronically varied by a special circuit known as a reactance tube and the other where the variation is brought about by electromechanical means. Almost all sweep generators today are a-c operated and make use of the 60-cps source from their own power-supply circuits as the modulating signal.

In the electromechanical systems, the 60-cps signal is applied to some form of vibrating mechanism in such a manner that the strength of the 60-cps signal will control the amplitude of vibration and the 60-cps frequency controls the rate of vibration. This special vibrating mechanism is mechanically ganged either to the rotor plates of a special type of variable capacitor which effectively varies the capacitance of the oscillator tank circuit, or, more frequently, to a special metal plate mounted near the inductance of the oscillator tank circuit. The vibration of this plate varies the effective inductance of the tank circuit. The greater the 60-cps voltage, the greater will be the frequency deviation. Consequently, regulation of the amount of 60-cps modulating voltage will control the amount of frequency deviation to the quantity desired.

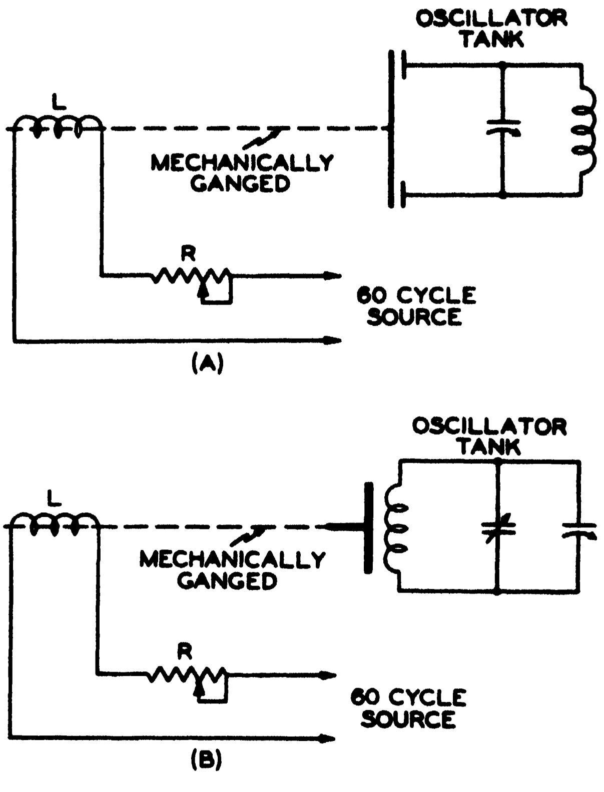

Fig. 16-9.—In the schematics of the modern mechanical type of sweep oscillator shown here, (A) indicates capacitive variation and (B) inductive variation.

The two drawings of Fig. 16-9 illustrate these two basic mechanical systems; part (A) represents the case where the capacitance is varied and part (B) the system where the inductance is changed. In both circuits the inductance *L* represents the vibrating mechanism and the potentiometer *R* controls the amount of 60-cps modulating signal applied to inductance *L*. Potentiometer *R* is often called the deviation control or ***sweep width control***. The oscillator tank can be part of most types of r-f oscillator circuits. The dashed line in each drawing indicates the mechanical ganging between the coil receiving the 60-cps voltage and the movable tank-circuit component. In part (B) the heavy vertical line near the tank-circuit inductance represents the movable plate.

With the electronic type of sweep generator the 60-cps modulating signal has the same effect upon the reactance-tube circuit as in the mechanical units. A simplified reactance-tube circuit showing the application of the 60-cps modulating voltage appears in Fig. 16-10. This circuit operates by placing across the os-

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[2] See Chapter 14 for a more detailed comparison of modern sweep generators.

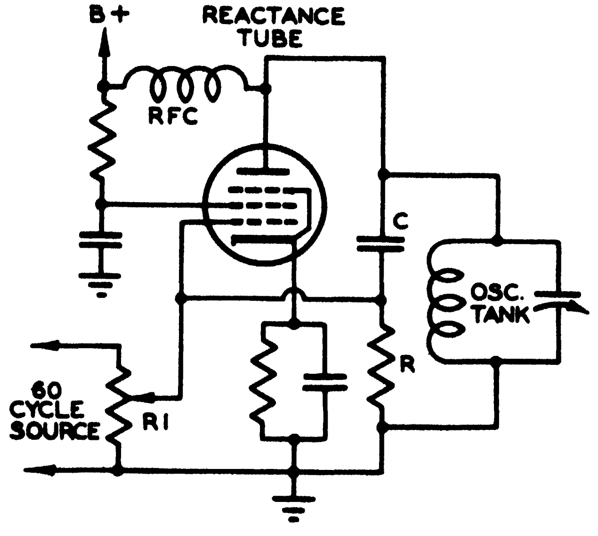


Fig. 16-10.—A simplified reactance-tube circuit.

cillator tank circuit an impedance which varies in accordance with the changes in the transconductance of the tube. According to the circuit components at the output of the tube, the reactive part of this impedance can be either inductive or resistive. For the circuit under discussion, capacitance *C* and resistance *R* are the important elements and they cause the impedance to be capacitive. The resistive part of the impedance is made to have negligible effect by making the reactance of *C* more than five times greater than the resistance of *R* and also by using a tube having a high plate resistance.

In such circuits the plate current is always in quadrature (i.e. 90° out of phase) with the voltage across the series *C* and *R* network (the output voltage), with the current leading the voltage. The injected capacitance for this circuit is directly proportional to the transconductance or g_m of the tube. Thus, an alternating signal applied to the grid of the tube will change the g_m of the tube an equal amount on either side of its steady-state value. This change in g_m will, in turn, vary the injected capacitance and hence cause frequency modulation of the oscillator. The applied 60-cps signal to the grid of the tube accomplishes this effect. The stronger the 60-cps voltage, the greater the change in g_m, the greater the injected capacitance variation, and hence the greater the resulting frequency deviation of the f-m signal. In the circuit of Fig. 16-10 potentiometer *R1* is the deviation or sweep width control.

From this analysis, in conjunction with what was previously said about bandwidth it should be understood that for visual alignment the sweep generator must have an f-m output whose center frequency is approximately equal to the center frequency of the response curve being observed and whose minimum peak-to-peak frequency deviation should be ***more than*** capable of covering the bandwidth of the circuit. Since the center frequency and bandwidth of the tuned circuits in a-m, f-m, and television receivers differs so much, the proper choice of the sweep generator to be used is quite important.

Sweep-Generator Controls and Terminals

There are many different types of modern alignment sweep generators on the market today. Basically they are primarily designed to furnish the necessary signals for visual alignment. However, a great deal of these sweep-generator units also include a marker generator. This was indicated at the beginning of the chapter. Most of the controls and terminals on sweep-generator units are all designed to perform the same function but the names and placement of these controls vary quite a lot. This is analogous to the situation relative to the terminology of the controls on oscilloscopes as indicated in Chapter 10. In this section we will discuss those controls and terminals that are pertinent to visual alignment. To try and discuss all the controls on the different sweep generators is beyond the scope of this book.

There are four major types of controls and two terminal connections existing on most sweep generators that are important to proper alignment. The function of these four major controls are as follows:

1. Selection of the center frequency of the f-m output signal.
2. Regulation of the amplitude of the f-m output signal.
3. Determination of the amount of frequency deviation of the f-m output signal.
4. Control of the phase between the signals used for modulating the f-m wave and that used for the horizontal-deflection voltage of the oscilloscope.

The two principal terminals are as follows:

1. The terminal where the f-m output signal is taken off.
2. The terminal where the modulating signal can be taken off to feed the horizontal input terminals of the oscilloscope.

Center frequency selection of the f-m signal may be accomplished by a single continuous control alone, or by a range switch in conjunction with a continuous vernier control which finally selects the exact frequency. The latter type of arrangement is more common because it enables the sweep generator to cover a wide frequency range. The regulation of the amplitude of the f-m output signal may be accomplished by a single continuous control or by a step attenuator switch in conjunction with a continuous vernier control. The other two types of controls are usually accomplished by a single knob. Both types of terminals referred to may be of the cable type of connection or just a simple binding-post connection. The former type of connection is used a great deal for the f-m output terminals. The positions of these controls and terminals on present-day sweep generators are so varied that it would be difficult to draw a representative diagram illustrating typical locations.

In order to indicate the various names that are given to these controls and terminals we have listed in Table I the names *exactly* as found on quite a few of the modern sweep generators.

TABLE I

A. Center Frequency Selection

Selects Exact Frequency	*Selects Frequency Range*
Frequency Adjustment	Range Selector
Tuning	Selector
Vernier	Frequency
Fine Frequency Adj.	Coarse Frequency Adj.
Sweep Frequency	Sweep Frequency Band Selector
——	Range
——	Band

B. Amplitude Control of F-M Signal Output

Fine Attenuation	*Step Attenuator Switch*
Output Control	Output Multiplier
Fine Attenuation	Step Attenuator
Fine Attenuator	Attenuator
Fine Attenuation	Coarse Attenuation
R.F. Amplitude	R.F. Attenuator
R.F.	——
Microvolts	——
Attenuation	——
Output	——

C. Control of Amount of Frequency Deviation

Sweep Width
Sweep Mc.
Megacycle Sweep
Dev. Mc.

D. Control of Phase

Phase
Phasing
Phasing Control
Phase Shifter
Scope Phase

E. F-M Signal Output Terminal

Output
R.F. Output
R.F. Out
R.F.

F. Terminal for Horizontal-Deflection Voltage

Horizontal Sweep
Horizontal Sync
Scope
Sweep Out
Phasing
AF Out (Sync)
60~

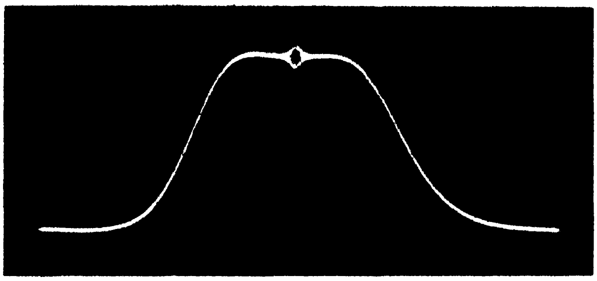
In some generators, although there is a control which selects the exact center frequency of the f-m wave, there is no name designated to it. This is indicated in section A of Table I. In section B, the last four names indicate that there is no attenuator switch for coarse control of the f-m signal amplitude.

Marker Generation

It is often valuable when employing visual alignment to be able to identify certain frequency points on the response curve or the frequency limits of the curve. Today two principal methods are employed whereby such frequency points are identified. Whatever method is used, the points are indicated by changes in amplitude on the response curve at specific frequencies.

Both methods of frequency identification were mentioned in Chapter 14. One system uses an absorption-type wavemeter, which, by virtue of its energy-absorbing qualities produces a decrease in amplitude (i.e., a dip) at the point in the response curve whose frequency is the same as the resonant frequency of the wavemeter. The other system makes use of an unmodulated r-f signal within the pass-band of the circuit under test. This r-f signal is used in conjunction with the f-m sweep signal. The use of such a signal as this produces a change in amplitude above *and* below that of the response curve.

Whatever system is used, that part of the curve in which there is this change in amplitude is generally referred to as a *marker point* or just a *marker.* Today the method using an unmodulated r-f signal is employed most often. Such r-f signals are obtainable from special *marker generators* which are usually crystal-controlled. These marker-generator systems may be a part of the sweep generator or a separate unit. In Fig. 16-11 appears a typical over-all television i-f response curve with a 44.25-Mc marker indication in the center of the curve. The unmodulated r-f signal used is commonly known as the *marker signal* or *marker frequency.*



Courtesy Zenith Radio Corp.

Fig. 16-11.—A typical over-all video i-f response curve having a marker signal in the center of the curve.

When the marker generator is a separate unit, its signal output can be injected anywhere in the receiver that is desired. There are two types of combined sweep and marker-generator units. In one type the marker signal output is available from a separate terminal and hence can also be applied to any desired part of the receiver circuit. In the other type, the marker signal is available together with the f-m output signal of the generator. Hence, in this latter type of unit, the marker signal is automatically injected at the same point in the receiver to which the f-m signal is fed. Of course, it is possible, in this latter type of unit, to use *either* the marker signal or f.m. alone without the other.

In many of the marker generators, whether separate units or parts of the sweep generators, a variable frequency oscillator is used to determine the marker frequency. Most of these generators employ a crystal oscillator circuit for accuracy of calibration of the r-f marker signal, with the usual provision for external connection of the crystal. In some units, the marker signal is simply obtained from the harmonics of certain crystal oscillators. In this case, the marker frequency is usually not continuously variable.

The trend today is to incorporate the marker generator as part of the sweep generator. Each one of these units has two major controls, one which selects the marker frequency and the other which controls the amplitude of the marker signal. Most of these units have a special plug connection for inserting a crystal. In those units where the marker signal is available separately from the sweep generator, there is also an extra terminal for taking out this marker signal.

The names given to these principal controls and terminals are not as varied as are those of the sweep generator. For example, for marker amplitude control, some of the names usually given are *marker output, marker, marker att.* and *marker injection.* For the marker-frequency control, the most common names are *marker* and *marker frequency.* For the crystal socket, the terminology used is primarily *crystal* and in some cases just *XL,* which is an abbreviation for the word crystal. For those units that have provision for separate marker output terminals, these terminals are termed just *marker* and *external marker.*

FORMATION OF THE PATTERN

Now that we have the fundamental information about the equipment needed for visual alignment and the facts about the bandwidths of the different receivers' tuned circuits we are now ready to discuss the formation of the pattern on the oscilloscope. In this respect we have to know something about the signal inputs to the oscilloscope and where they are obtained before the actual process of pattern formation is analyzed.

Rectification of the Tuned Circuits' Output

At the very beginning of this chapter we mentioned that rectification of the signal output from the tuned circuit or system under alignment is necessary for proper visual observation of the response curve.

Examination of receiver manufacturers' service data relative to visual alignment or that found in servicing text books will indicate that the vertical-deflection voltage for the oscilloscope is generally secured from the detector circuit of the receiver in question. There may be some variations as to exactly the part of the circuit this voltage is secured from. This depends upon the type of receiver that is to be aligned. Whatever the case may be, these references indicate that *rectification* of the signal output from the tuned circuit or system is necessary in order to develop the resonance curve on the screen of the cathode-ray tube. Let us analyze this problem of rectification and find out why it is necessary.

First of all it is important that we remember that an f-m signal input is necessary in order to cover the bandwidth of the tuned signal. Let us assume that an f-m

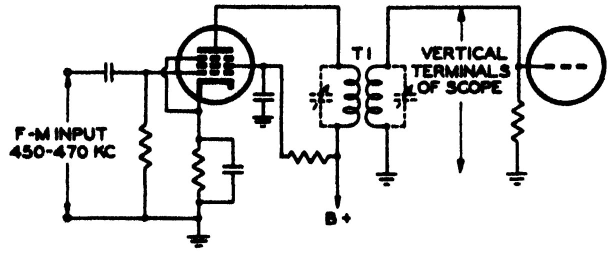


Fig. 16-12.—Circuit hookup for examining the output of an i-f amplifier system on an oscilloscope when using an f-m signal input.

signal with a center frequency of 460 kc and a peak deviation of 10 kc is being fed into an i-f amplifier and also that the vertical terminals of the oscilloscope are connected across the secondary of this transformer, all as indicated in Fig. 16-12. If the transformer is untuned and does not offer any frequency discrimination between 450 kc and 470 kc then the pattern that will appear upon the oscilloscope will be just an f-m wave of constant amplitude containing the same frequency deviation as the input f-m signal. (The frequency response of the vertical amplifiers must be flat to at least 500 kc in order to reproduce this signal.)

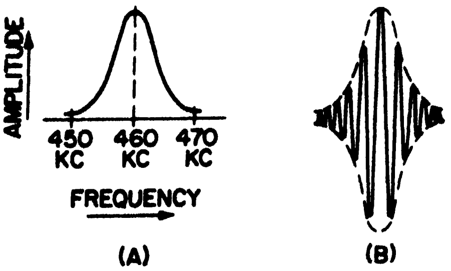


Fig. 16-13.—Part (A) represents the single-peaked response of tuned transformer *T1* in Fig. 16-12 and part (B) is the pattern that would appear upon the oscilloscope.

Now let us assume that *T1* is a double-tuned transformer, as indicated by the dotted trimmer capacitors across the coils, and that the circuit is resonant at 460 kc. The graphical resonance curve for this tuned transformer is illustrated in Fig. 16-13A. Note the frequency limits of this curve. With the same f-m signal input as before and the tuned circuit arrangement as indicated, a pattern will appear on the oscilloscope but it will *not* be the simple resonance curve of Fig. 16-13A. The pattern which will appear will be an f-m wave that is amplitude-modulated with the *envelope* of the resonance curve. This is indicated in Fig. 16-13B. The reason why such a pattern appears on the oscilloscope is explained as follows.

The tuned circuit has a certain degree of selectivity as seen by its response curve shown in Fig. 16-13A and hence imparts amplitude-modulation effects to the f-m wave passing through it. The envelope of the resonance curve appears above *and* below the pattern of Fig. 16-13B because it changes the amplitude of the input f-m signal above and below its zero reference axis. In other words the selectivity of the tuned circuit affects both half-cycles of each cycle of the f-m signal in the same manner.

Although from the pattern of Fig. 16-13B we can tell the shape of the resonance curve of the tuned circuit from either the upper or lower envelope, such a method is not very practical for a number of reasons. For exact interpretation of alignment patterns, the actual resonance curve is desired since it is less confusing. Another important reason is that many of the service oscilloscopes in use do not have a frequency range adequate to pass frequencies as high as 500 kc. Therefore, it is only possible to visually align a-m receivers by the method of Fig. 16-12 if the oscilloscope has a frequency range well beyond the i.f.'s of the receiver. For f-m and television receivers, the situation is much different. The i.f.'s of the majority of f-m receivers are 10.7 Mc and for the video circuits of television receivers the i.f.'s are generally above 20 Mc. There are a few, but very expensive, laboratory oscilloscopes that have a frequency response to 10 Mc and higher.

If the signal output from the tuned transformer circuit is rectified before it is applied to the oscilloscope, it is possible to use almost any type of oscilloscope for visually observing the resonance curve of the circuit. The new circuit is shown in Fig. 16-14. This circuit is very similar to the diode detector circuit of a-m re-

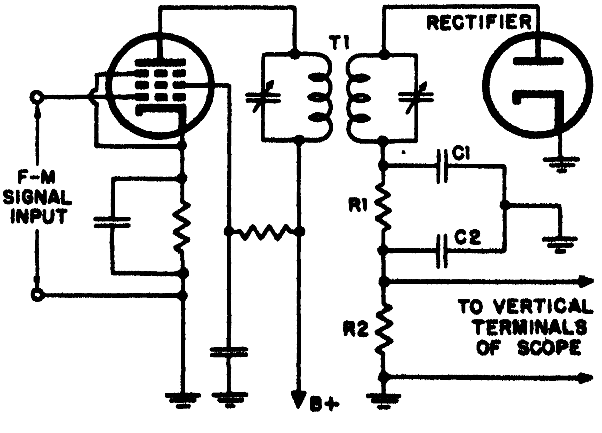


Fig. 16-14.—Schematic showing how the f-m signal output from the tuned circuit is rectified before being applied to the oscilloscope.

ceivers or the video detector circuit of television receivers. The same amplitude-modulated pattern of Fig. 16-13B appears in the output of *T1*. This pattern appears to the diode as a simple a-m wave, and, consequently, the signal becomes rectified *and* detected by this tube. Since a simple diode circuit cannot respond to changes in frequency, it does not detect the f-m part of the signal. Consequently, due to rectification, exactly half the pattern of Fig. 16-13B results. Components *R1, C1,* and *C2* of Fig. 16-14 serve as an r-f filter for the f-m part of the signal and hence only the envelope of the signal appears across the diode load, resistor *R2*. Since the envelope of the signal is repeated at the same rate as the modulating frequency of the f-m signal, the frequency response of the oscilloscope need only be so great as this modulating frequency. This holds true no matter what the i.f. of the receiver being aligned is. In practice this modulating frequency is always within the audio range, being 60 cps in the majority of cases. We can see that by this latter method of visual alignment almost any type of oscilloscope can be used as long as the lower limit of its vertical frequency response is below the modulating frequency of sweep generator, and this is true of most oscilloscopes.

Detector Probe

When the detector of the receiver is used, it is not possible to observe the resonance curve of *each* individual i-f stage. The only single i-f resonance curve that it is possible to see is that of the last i-f stage of the receiver (i.e. that preceding the detector), because this is the only stage whose input signal can be secured from the sweep generator and yet have its output signal fed directly to the detector. When other i-f stages are aligned, the resonance curve seen on the oscilloscope is actually a *combination* of the individual resonance curves of the i-f stages *between* the sweep generator and the detector circuit of the receiver.

Observation of individual resonance curves is possible by using an auxiliary detector for rectification of the output signal rather than the detector of the receiver. In this manner the auxiliary detector can be moved to whatever stage is to be aligned and only the resonance curve of that stage will appear on the oscilloscope. Alignment through the observation of individual resonance curves is often desired in television receivers where the i-f stages employed may each have different resonance frequencies and/or different types of coupling. In actual single-stage resonance alignment a system is used where the complete circuit is enclosed in a probe for ease of manipulation. This is generally referred to as a *detector probe* or *traveling detector probe.*

The circuit arrangement of a typical detector probe is shown in Fig. 16-15 where a germanium crystal type 1N34 functions as the detector. In this circuit the positive side of the crystal refers to the anode or plate of the detector and the negative side indicates the cathode.

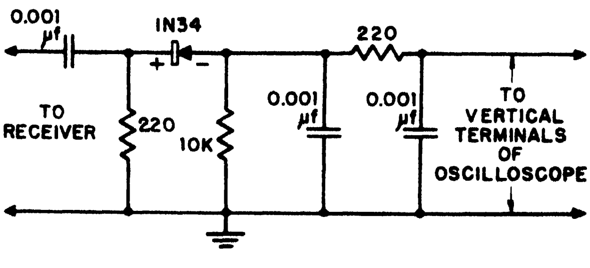


Fig. 16-15.—A typical detector probe circuit.

Using a unit such as this requires no heater voltages and hence it is an easy matter to enclose the complete circuit in a probe. The detector circuit of Fig. 16-15 is representative of the type as used in stage-by-stage alignment of television receivers. The values of the components may vary somewhat from those indicated in the drawing. The two 0.001 μf capacitors and the 220-ohm resistor combination at the right side of the probe (the end connected to the vertical input terminals of the oscilloscope) functions as a filter to the r-f signals. The 10,000-ohm resistor acts as the load for the detector. The left side of the detector circuit is connected to the output circuit of whatever stage is to be aligned.

The manner in which the positive and negative terminals of the crystal detector are connected and the number of vertical amplifiers used in the oscilloscope helps determine whether the response curve appears upright or inverted. To change an inverted resonance curve simply means reversing the connections of the crystal detector.

The Horizontal-Deflection Voltage

One of the most important aspects of the visual-alignment procedure is the use of the correct signal for the horizontal-deflection voltage of the oscilloscope. In order to reproduce the resonance curve on the oscilloscope the signal used as the modulating voltage for the f-m wave inside the sweep generator should also be used as the horizontal-deflection voltage for the oscilloscope. In other words, if the modulating signal of the f-m wave varies linearly with respect to time, the horizontal-deflection voltage should also vary linearly. In like manner if the modulating signal of the f-m wave is sinusoidal, the horizontal-deflection voltage must be sinusoidal.

One common error that is usually made in visual-alignment procedures is that the sawtooth time base of the oscilloscope is often used as the horizontal-deflection voltage *regardless* of what type of signal modulates the carrier inside the sweep generator. The resonance curve produced by this method is not a true curve displayed on a linear scale but is distorted in some fashion. Since we are used to graphically representing resonance curves on a linear scale, it is important that we try to reproduce such curves upon the oscilloscope.

The majority of modern sweep generators employ sine-wave modulation so that a sine wave should also be used as the horizontal-deflection voltage for the oscilloscope. As previously pointed out most sweep generators have an output terminal where this sine-wave modulating signal could be taken off and fed directly to the horizontal input terminals of the oscilloscope. Practically all oscilloscopes have some arrangement on the front panel whereby the internal sawtooth generator of the unit is switched out of the horizontal amplifier circuit and the horizontal input terminals switched in.

Many oscilloscopes also have means whereby the 60-cps signal from the power-supply circuit of the unit may be used as the horizontal-deflection voltage instead of the sawtooth time base. Since the sine-wave modulating signal of the majority of sweep generators is also obtained from its own power-supply circuit, it is, therefore, possible to utilize the internal 60-cps sine-wave sweep of the oscilloscope as the horizontal-deflection voltage instead of connecting the 60-cps output from the sweep generator to the horizontal input terminals of the oscilloscope. It is suggested that the 60-cps sine wave from the sweep generator be used because in that manner phase control between the horizontal-deflection voltage and the modulating signal of the f-m output of the sweep generator is possible.

If the frequency of the horizontal-deflection voltage is the same as, or an integral multiple of, the modulating signal of the f-m wave, synchronization of the pattern on the oscilloscope will be complete. However, in order to reproduce a single pattern upon the screen, the frequency of both these signals should be the same. This, of course, is assured by using the same modulating signal from the sweep generator for the horizontal-deflection voltage.

Tracing Out the Curve

In this section, we will explain how the response of an undercoupled i-f transformer is traced out on the screen of an oscilloscope, first using a modern sweep generator employing sine-wave modulation, and then using a wobbulator employing a rotating capacitor. The reason we are including the wobbulator method is to give a comparison between what has been used in the past and what is commonly employed today.

Using the Wobbulator

We mentioned previously that for the reproduction of the resonance curve on a linear scale the modulating signal of the f-m wave and that causing horizontal deflection should have the same shape. When a wobbulator is employed, we do not deal with an actual modulating signal as in the modern sweep generator, but rather with a rotating capacitor. In the former method, the amplitude of the sine wave determines the amount of frequency deviation, whereas in the latter case it is the amount of capacitance variation of the rotating capacitor which determines the amount of frequency deviation. It is the curve of frequency versus time produced by this rotating capacitor which is analogous to the sine-wave modulating signal of the other type of unit. The shape of this frequency versus time characteristic, which is standard, is shown in Fig. 16-16. Between points *a* and *g* this curve represents three revolutions of the rotating capacitor.

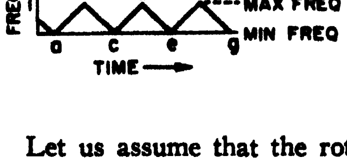


Fig. 16-16.—Triangular frequency versus time characteristic of the rotating capacitor used in wobbulators.

Let us assume that the rotating capacitor is connected across the parallel tuned tank circuit of an oscillator. This means that when the capacitance of the rotating capacitor is a maximum (i.e. plates fully meshed) the instantaneous frequency of oscillation is a minimum and when the capacitance is a minimum (i.e. plates completely open) the instantaneous frequency is a maximum. Thus, minimum frequency points *a*, *c*, *e*, and *g* of Fig. 16-16 occur when the plates of the capacitor are fully meshed and maximum frequency points *b*, *d*, and *f* occur when the plates are completely open.

Since the waveshape is linear, the linear sawtooth time base of the oscilloscope can be used for the horizontal-deflection voltage. In order to understand how a pattern is traced out let us assume a starting condition where the instantaneous frequency of the wobbulator is a minimum and where the sawtooth voltage of the oscilloscope starts increasing from its minimum value. The frequency of rotation of the capacitor and that of the sawtooth sweep are made equal so that it takes the same amount of time to complete one triangular wave as it does to complete one cycle of sawtooth. Let us further assume that the wobbulator input is fed to a loosely-coupled i-f system, where the output signal after detection is a single-peaked resonance curve.

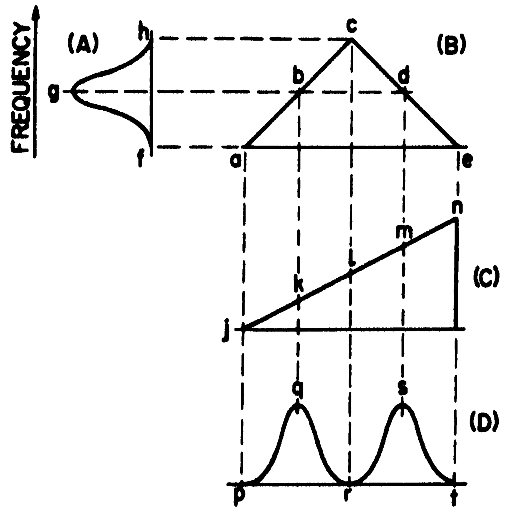


Fig. 16-17.—Graphical analysis of how a single-peaked symmetrical resonance curve is reproduced using a wobbulator.

Refer to Fig. 16-17 (The notation here is distinct from Fig. 16-16.) Part (A) represents the single-peaked resonance curve (shown in a side position for the sake of illustration), part (B) the triangular wave, and part (C) the sawtooth signal. Since one revolution of the capacitor causes an indicated change in frequency from minimum to maximum and back to minimum again, the f-m signal sweeps over the tuned circuit *twice.*

This means that when curve (B) in Fig. 16-17 goes from *a* to *c* the f-m signal sweeps over curve (A) from *f* to *h* and when curve (B) continues from *c* to *e,* the f-m signal sweeps *back* over curve (A) from *h* to *f.* Let us now see what this means when the detected output signal is fed to the vertical terminals of the oscilloscope.

The sawtooth signal of Fig. 16-17C, when acting alone just causes the beam to be deflected along the horizontal, starting from the extreme left of the screen, which corresponds to point *j* on the sawtooth. With the input signal applied to the vertical-deflection terminals and the sawtooth signal to the horizontal-deflection amplifiers, the combined action is such that during the first quarter of a cycle of sawtooth signal, that is, from points *j* to *k,* the resonance curve is traced out from *f* to *g.* This is indicated in part (D) from *p* to *q.* During the next quarter of a cycle of sawtooth signal, from time *k* to *l,* the second half of the resonance curve of Fig. 16-17A is traced out from points *g* to *h.* This is indicated in Fig. 16-17D from points *q* to *r.*

Thus far we see that during one half-cycle of sawtooth signal, the rotor of the capacitor has turned only half a revolution and the resonance curve has been traced out once. For the next half-cycle of sawtooth signal, the capacitor completes its revolution and the resonance curve is swept again and a second resonance curve appears on the screen. This latter curve is indicated in Fig. 16-17D during time *r* to *t.* This second resonance curve tracing from *r* to *t* corresponds to time *h* to *f* of part (A) of the drawing. This correspondence is difficult to detect from the symmetrical single-peaked patterns shown in the illustration but if the curves were not symmetrical it would be easy to see. The drawing of Fig. 16-18 illustrates such a case. This figure is analogous to Fig. 16-17. From examination of part (D) of the latter figure one can readily see that the two resultant resonance curves are traced out in *opposite* directions. In other words, the two resonance curves of Fig. 16-18D are mirror images of each other.

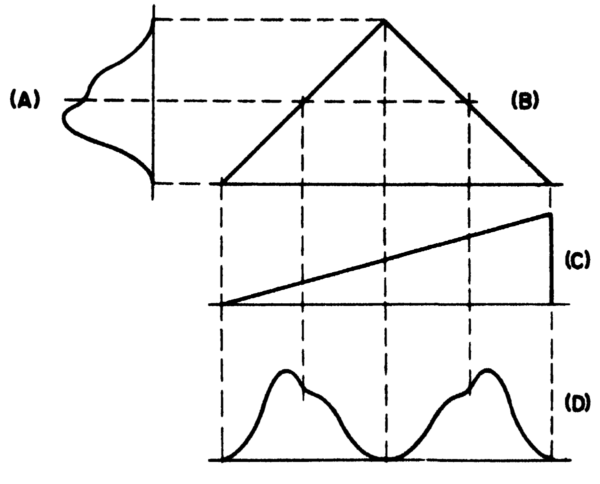


Fig. 16-18.—Graphical analysis of how an asymmetrical resonance curve is reproduced using a wobbulator.

In Fig. 16-19 appear two oscillograms of resonance curves produced by a wobbulator. The double pattern of photo (A) is that of a symmetrical single-peaked

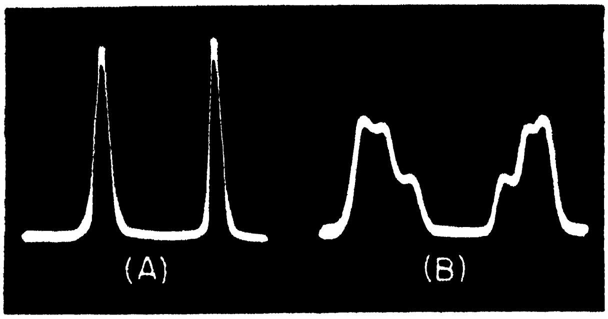


Fig. 16-19.—Oscillograms of resonance curves obtained with a wobbulator.

resonance curve whereas photo (B) is a double pattern of an unsymmetrical resonance curve which is seen to contain three peaks. These two photos corroborate the previous theory about the manner in which the f-m signal traces out the resonance curve.

In the previous analysis, the time that it took the capacitor rotor to complete one revolution and the sawtooth sweep signal to complete one cycle were equal. This is the same thing as saying that one cycle of sawtooth is completed during one cycle of the triangular wave. How would the pattern appear if the frequency of the sawtooth were twice that of the triangular wave? In such a case the complete resonance curve is traced through twice, as in the other method, but appears differently on the screen.

Let us refer to Fig. 16-20 and see how an unsymmetrical resonance curve, part (A), is traced out on the screen under these conditions. Part (B) of this drawing is the rotating capacitor's frequency versus time characteristic and part (C) is the sawtooth time base of the oscilloscope. Since two cycles of sawtooth signal are employed, the spot, after tracing out one resonance curve, returns to its starting position to trace out the curve once more. During the first half-cycle of capacitor rotation, or from time *a* to *b* on curve (B), the complete resonance curve of part (A) is swept from *d* to *e*. During this time the first cycle of sawtooth wave of part (C) is active, which means that the complete resonance curve appears on the screen. This is indicated in part (D) of the drawing by the solid curve traced out from *j* to *k*. During the latter half-cycle of capacitor rotation, the triangular wave of (B) goes from time *b* to *c*, which means that the f-m signal sweeps across the tuned circuit again but from *e* to *d* on the resonance curve of part (A). Since the second cycle of sawtooth starts out from the same spot as the first cycle, the resonance curve is traced out again but this time in the opposite sense. This means that the curve that results during this latter cycle of sawtooth operation is traced out from *j* to *k* as indicated by the *dashed* curve in part (D) of the drawing.

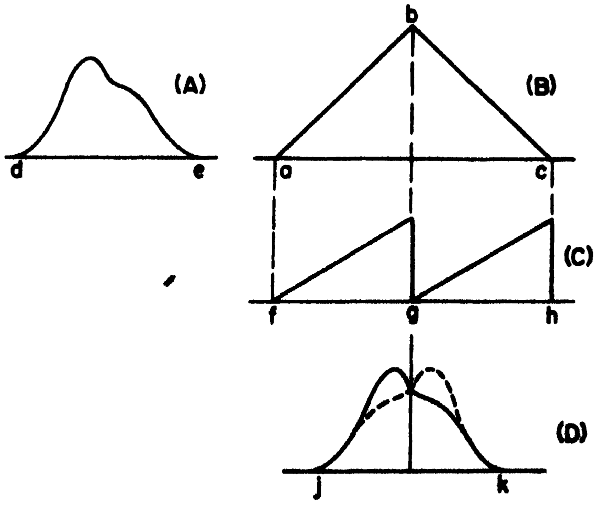


Fig. 16-20.—Graphical analyses of how an unsymmetrical resonance curve will be produced when the frequency of the sawtooth deflection voltage is twice that of the triangular wave.

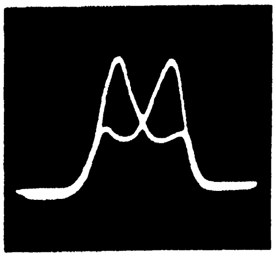


Fig. 16-21. — Oscillogram of an unsymmetrical resonance curve proving the analyses of Fig. 16-20.

The photograph of Fig. 16-21 is a typical oscillogram proving the above analysis. From this picture the response in question is seen to be a double-peaked unsymmetrical curve. If the resonance curve that is to be traced out by this latter method is symmetrical, then only one trace would appear upon the screen. This is so because both curves when reproduced upon the screen coincide with each other. This is shown in the two oscillograms of Fig. 16-22. The photograph (A) is that of a symmetrical single-peaked curve and (B) is a symmetrical double-peaked curve.

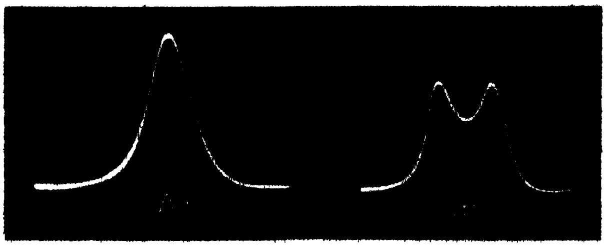


Fig. 16-22.—When the curves are symmetrical, only one curve will appear on the screen as indicated by the single-peaked curve of (A) and the double-peaked curve of (B).

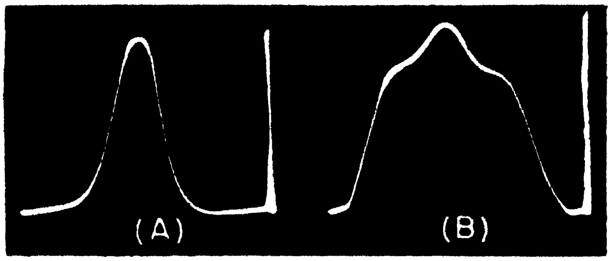


Fig. 16-23.—If the frequency of the sawtooth is twice the triangular wave and if the second cycle of sawtooth is made inactive, only single images would result but vertical lines would also appear as indicated here.

If the second cycle of sawtooth sweep of Fig. 16-20C can be made inactive then only one resonance curve will appear upon the screen. The sawtooth signal will, therefore, only be active during the first half-cycle of capacitor rotation. Frequency modulation of the oscillator is taking place during the complete cycle of the rotating capacitor but each spot position is related to only one frequency and no matter how asymmetrical the pattern is, only a single trace appears. However, this trace is also accompanied by a vertical line, which represents the vertical displacement during that portion of the frequency-modulating cycle when there is no horizontal displacement. The two photos of Fig. 16-23 illustrate this latter method. (A) is a symmetrical single-peaked curve while (B) is a triple-peaked curve that is slightly unsymmetrical. Note the vertical line on the right side of each oscillogram.

Using A Modern Sweep Generator

Let us now see how the resonance curve is traced out upon the screen of the oscilloscope when using a sweep generator that employs sine-wave modulation of the f-m signal. As mentioned before, the horizontal-deflection voltage should also be a sine wave and of the same frequency as that of the modulating signal in the sweep generator. With this type of system, the resonance curve is also traced out twice but the tracing is accomplished in such a manner that the two resonance curves always coincide with each other no matter what the shape of the curve. This, of course, is predicated upon the supposition that the sine-wave modulating signal and sine-wave horizontal-deflection signal are in phase.

Let us assume that we are to observe an unsymmetrical resonance curve by the sweep-generator method, that the vertical input terminals of the oscilloscope are connected to a detector circuit and the sweep generator feeds in the proper f-m signal to the tuned circuit being tested. The sine-wave modulating signal of the generator is also fed directly to the horizontal input terminals of the oscilloscope (the internal sawtooth sweep of the oscilloscope is not being used). In order to best understand how the pattern is traced, we will study some basic actions first.

First of all it should be remembered that it is the amplitude of the modulating signal that determines the amount of frequency deviation of the f-m signal inside the sweep generator. Thus, at the positive peaks of the modulating sine wave the instantaneous frequency of the f-m signal is a maximum and at the negative peaks it is a minimum. In the following analysis we will consider the sine wave as starting out at the negative peak.

In Fig. 16-24 part (A) represents one cycle of this sine wave and part (B) is the unsymmetrical resonance curve of the tuned circuit. During the time that the sine wave at (A) increases from a negative maximum at *a* to a positive maximum at *b* the frequency of the f-m wave increases from a minimum to a maximum, which means that the resonance curve of (B) is swept through once, from time *d* to *e*. During the latter half-cycle of the sine wave at (A), from time *b* to *c*, the resonance curve is traced out again, but this time the curve is traced from time *e* to *d* because the f-m signal is decreasing in frequency from a maximum to a minimum.

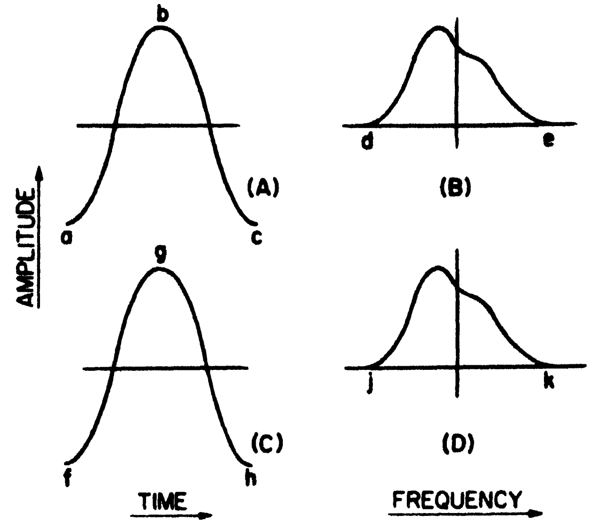


Fig. 16-24.—Graphical analysis of how an unsymmetrical resonance curve is produced using a modern sweep generator.

The sine wave at (C) in Fig. 16-24 represents the horizontal sweep voltage of the oscilloscope and the pattern at (D) the reproduced resonance curve. Note that it is in phase with, and equal in frequency to, that of part (A). From time *f* to *g*, the sine wave at (C) is increasing in amplitude from a maximum negative voltage to a maximum positive voltage. Considering the action of this part of the signal alone and with no signal on the vertical plates, the spot would trace a horizontal line across the screen from left to right. With the vertical input signal applied, which is the resonance curve of part (B), the action of both deflection fields upon the beam is such that it causes the spot to trace out the resonance curve on the screen from *j* to *k*, as seen in part (D) of the drawing.

Since the horizontal sweep voltage is sinusoidal, when considering the action of this signal at (C) alone, the spot sweeps back across the screen from right to left during time *g* to *h*. During this time, the f-m signal is also sweeping across the resonance curve of part (B) from right to left, that is, from time *e* to *d*. Consequently, when both signals (B) and (C) are active on the beam, the spot traces out the resonance curve on the screen from *k* to *j* of part (D). In this case the second resonance curve will exactly coincide with the first and only one pattern will be seen.

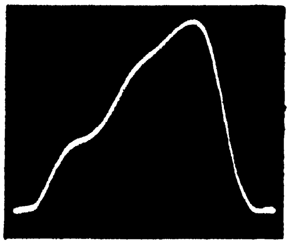


Fig. 16-25.—An oscillogram of an unsymmetrical resonance curve produced by a modern sweep generator.

From the above analysis, one can readily see how much simpler it is to observe a resonance curve by this method than by the wobbulator method. In Fig. 16-25 is an oscillogram of an unsymmetrical resonance curve that has been obtained by the use of the method just outlined. Note that only one trace appears upon the screen.

Phase Control in Visual Alignment

In the analysis just completed, the sine-wave modulating signal of the sweep generator was considered to be in phase with the sine wave used as the horizontal-deflection voltage for the oscilloscope. As mentioned, this is necessary in order that the two resonance curves coincide.

By the time that the sine-wave sweep reaches the horizontal-deflection plates of the oscilloscope there will be numerous circuit components (resistors, capacitors, and inductors) that will cause phase changes in this signal. Consequently we see that it is almost impossible to have the two sine waves in phase with each other unless some special circuit is incorporated to remedy any undesired phase changes. When there is a phase difference between these two sine waves, instead of a single resonance curve appearing upon the screen, two will be seen. Both patterns will normally have the same shape. The amount of phase difference determines the separation between them. In effect one pattern appears to be the shadow of the other.

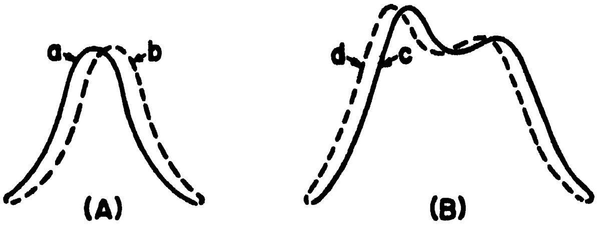


Fig. 16-26.—Double-image patterns that indicate the need for phase control.

In Fig. 16-26 we have illustrated two double-image drawings of resonance curves caused by the above condition. One trace of each double image is drawn dashed in order to distinguish it from the other. The drawing for part (A) is for a single-peaked resonance curve. For this particular case, the sine-wave modulating signal in the generator was leading the sine wave at the horizontal plates of the oscilloscope by a small amount. Solid-line curve *a* is traced out first and the dashed curve *b* is the return trace. Part (B) of Fig. 16-26 illustrates a double-image of a double-peaked resonance curve produced when the wave in the sweep generator was lagging the sine-wave deflection voltage of the oscilloscope. Curve *c* in this drawing is traced out first and dashed curve *d* is traced out on the return trace.

It is the special phasing control that is incorporated in most sweep generators that corrects for this difference in phase and makes the two curves coincide. The phasing control circuits used in the different sweep generators are usually *R-C* circuits. (This can be seen from examination of the sweep-generator schematics of Figs. 14-33, 14-40, and 14-42 in Chapter 14.) In Fig. 16-27 is a simplified form of one of these *R-C* circuits. The 60-cps sinusoidal modulating voltage is usually obtained from the secondary of some filament transformer in the power supply of the sweep generator as shown in the drawing. The sine-wave deflection signal is normally obtained from the same network and is fed to the horizontal input terminals of the oscilloscope. The variable arm of the potentiometer is connected to the high side of the horizontal input terminals, and the ground connections between the units made. Let us

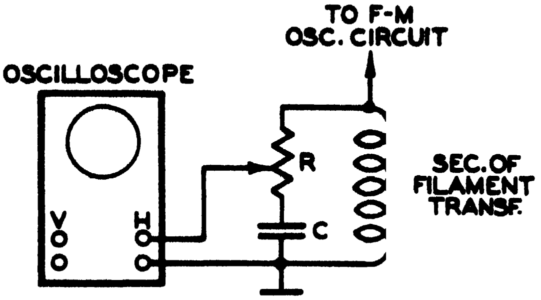


Fig. 16-27.—Simplified form of the *R-C* network used in sweep generators for phase control.

briefly analyze how this simple unit will make the two sine waves in phase and produce a single trace pattern.

Let us assume that the sine-wave modulating signal as obtained from the high side of the filament transformer has a certain phase angle due to the impedances in its circuit. The same source of sine-wave voltage is, as we already know, also used for the horizontal-deflection voltage of the oscilloscope. The impedance in the path of this deflection voltage is the sum of the impedances offered by the horizontal amplifier system of the oscilloscope plus the impedance of the combination of *C* and *R* in Fig. 16-27. This total impedance imparts a certain phase angle to the deflection voltage when applied to the horizontal-deflection plates of the oscilloscope.

It is seldom possible that these two circuit impedances, the one in the path of the generator's modulating sine wave and the one in the path of the oscilloscope's deflection sine wave (neglecting *C* and *R*), impart the same phase angle to their respective sine waves. This means that, in normal operation, the sine-wave modulating signal will not be in phase with the sine-wave deflection signal, and hence a double image will result. The phase angle produced by any impedance is dependent upon its ratio of reactance to resistance. Then if either of these can be changed, the phase angle can likewise be changed, as well as the total impedance. In the drawing of Fig. 16-27, resistor *R* is made variable so that the impedance and hence phase angle produced in the sine-wave deflection voltage can be varied.

In the process of alignment, potentiometer *R* of Fig. 16-27, which is the phasing control on the sweep generator, is adjusted until the double-image pattern on the screen appears as a single trace. When this occurs, both sine waves are in phase. Sometimes both images cannot be made to completely coincide. If this occurs, it is usually not due to any defect in the phase adjustment but to some other factor. The f-m signal output from the sweep generator may be distorted, the sine-wave modulating signal when finally applied to the f-m oscillator of the sweep generator may become distorted, or the horizontal amplifiers of the oscilloscope may distort the sine-wave deflection voltage. If variation of the phasing control cannot make the two images coincide at all, reversal of the resistance and capacitance as shown in Fig. 16-28 will usually correct this situation.

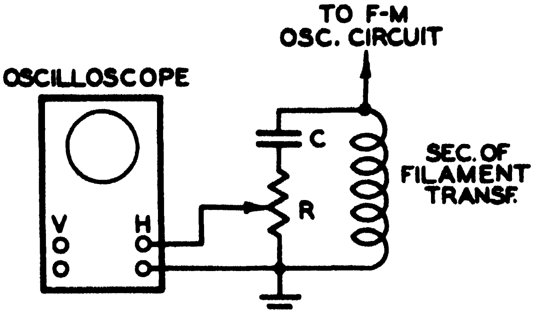


Fig. 16-28.—If the variation of *R* in Fig. 16-27 does not make the two images coincide, then reversal of *R* and *C*, as shown here, is necessary.

If the oscilloscope used obtains its horizontal sine-wave deflection voltage from a sweep generator that does not employ a phasing circuit, it is a simple matter to hook one up. The block diagram arrangement of this was indicated in Fig. 16-1C. The schematic arrangement is shown in Fig. 16-29. All that is required is a simple *R-C* phase-shifting network. The capacitor is connected across the horizontal input terminals of the oscilloscope, and the variable resistor is connected in series with the high side of the horizontal terminals. The other connections are as indicated in the drawing.

Since the amount of phase shift that may have to be corrected for will vary with the different instruments used and the circuits being aligned, the values of *C* and *R* are not always the same. Sometimes *R* may equal 100,000 ohms and *C* equals 0.0025 to 0.25 microfarads or *R* may equal 50,000 ohms and *C* equals 0.05 to 0.25 microfarads. If the connection illustrated in Fig. 16-29

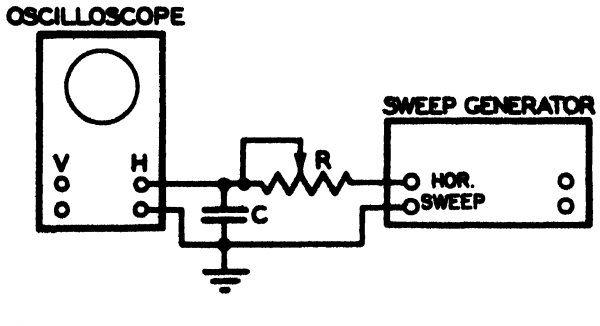


Fig. 16-29.—Method of hooking up a simple *R-C* phase-control network if the sweep generator does not employ one.

does not produce the correct direction of phase shift, then the resistor and capacitor should be interchanged.

Marker Indication

Now that we have a basic idea of how the pattern is traced out on the oscilloscope, we are ready to analyze how a marker indication can be made on a response curve. The production of a marker point in visual-alignment practice is based upon the principle of heterodyning. To understand properly how the marker signal is produced and the nature of the marker indications, it is best to review some of the basic principles involved in the process of heterodyning. This heterodyne operation is also referred to as mixing, frequency conversion, or beating.

Mixing Two Unmodulated R-F Signals

If two unmodulated r-f sine waves are mixed together in a heterodyne diode detector, the output from this detector will contain various signals. Among these are output signals having frequencies equal to those of the input waves and also those of the sum and difference frequencies of the two input signals. In most heterodyne diode detectors, it is the difference frequency, often referred to as the beat note, which is the one that is usually desired. If the frequencies of the two input signals in question are 1,000 and 1,050 kc, the resultant difference frequency will be equal to 50 kc and will have an amplitude equal to, or somewhat smaller than, that of the smaller amplitude input signal. If the output of this detector is fed into the vertical amplifier of an oscilloscope (using the internal sawtooth sweep of the oscilloscope as the horizontal-deflection voltage) a sine-wave pattern will be observed on the screen. A block-diagram arrangement for this is shown in Fig. 16-30.

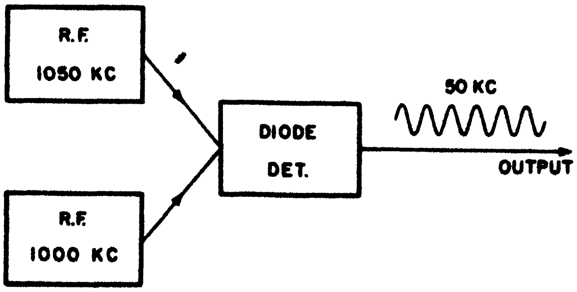


Fig. 16-30.—Block diagram showing how a heterodyne detector acts on two incoming unmodulated r-f signals.

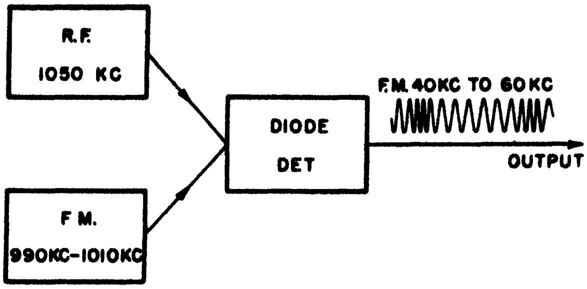


Fig. 16-31.—When an f-m and an unmodulated r-f signal are mixed in the diode detector, an f-m output signal will result.

Let us advance one step further in this heterodyne analysis and substitute an f-m signal, which is varying in frequency from 990 kc to 1,010 kc (a peak-to-peak frequency deviation of 20 kc), in place of the unmodulated 1,000-kc signal. The block-diagram arrangement for this appears in Fig. 16-31. When an f-m signal and an unmodulated r-f signal are mixed together in a heterodyne detector, the difference frequency signal output will always be an f-m signal. This resulting f-m signal will have the same total deviation as the input f-m signal, namely, 20 kc, but will have a center frequency equal to the difference between the unmodulated r-f signal and the center frequency of the input f-m signal. Thus, the resulting f-m signal output will be varying in frequency 10 kc on either side of a mean frequency of 50 kc or from 40 kc to 60 kc. When applied to the vertical input terminals of an oscilloscope the pattern that will appear will be f.m. The alternate spreading and bunching of the cycles of this signal will be seen if the frequency of the linear sweep of the oscilloscope is low enough (about 2 to 3 kc) to allow enough cycles to appear on the screen. Of course, the larger the screen size, the easier will be the examination of the f-m wave.

Mixing F-M and Unmodulated Signals

Let us now change the frequency of the unmodulated r-f signal from 1,050 kc to 1,000 kc but keep the f-m signal input at the same frequency and see what happens. Since we are again mixing an f-m and an unmodulated r-f signal, the resulting difference frequency signal output from the diode detector will likewise be an f-m wave. The frequency and general appearance of this output f-m signal is, however, somewhat different from that resulting from the previous mixing process. Since the frequency of the unmodulated signal is the same as the center frequency of the input f-m wave, the instantaneous frequency difference between the two signals at this point will be zero (i.e., there will be a zero beat), and for the extreme points of the resultant f-m wave, the frequency difference will be the same, namely 10 kc. In other words, as the input f-m wave swings from 990 kc to 1,000 kc to 1,010 kc, the resulting beat signal will be varying in frequency from 10 kc to 0 to 10 kc. This is indicated in the drawing of Fig. 16-32.

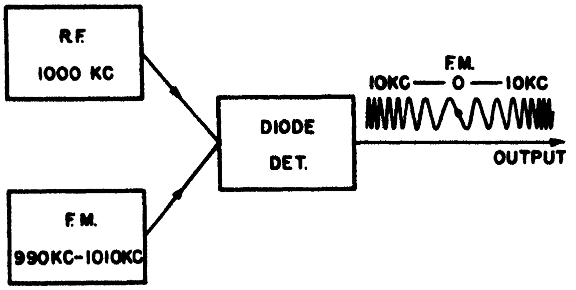


Fig. 16-32.—Beating an f-m and an unmodulated r-f signal to produce an f-m output with a zero center frequency.

This method of heterodyning is an interesting phenomenon in that during the process of mixing, at one particular point there will be no output from the detector. This occurs when the instantaneous frequency of the f-m input signal is equal to that of the unmodulated input signal. In the case under discussion, it occurs at the center frequency of the f-m signal. The resulting f-m signal output, an exaggerated drawing of which is shown in Fig. 16-32, is so arranged that it will be bunched at the ends and spread in the center. The degree of spread in this case is much wider than at the minimum frequency point for the f-m wave output of Fig. 16-31. This is so because the point of maximum spread for the situation of Fig. 16-32 is supposed to represent zero frequency output; that is, zero beat occurs at this point.

If this signal output were fed to an oscilloscope with a sawtooth horizontal sweep, it would appear as a continuous f-m wave. The point of zero frequency output is not perfectly defined. This point is indicated by a dot on the wave in the drawing of Fig. 16-32. The difference frequency heterodyne signal output in each of the above cases would, when using the sawtooth sweep of the oscilloscope, appear as indicated in the three previous drawings. Each wave would be approximately constant in amplitude. Let us now consider a change in the arrangement of Fig. 16-32 by using the sine-wave modulating signal of the sweep generator as the horizontal-deflection voltage for the oscilloscope. The new arrangement is illustrated in Fig 16-33A. The horizontal input selector switch of the oscilloscope is turned to the point where the sawtooth time-base generator of the unit is not used and the horizontal input terminals are connected directly to the input of the horizontal amplifiers. The output from the diode detector is still fed to the vertical amplifiers of the oscilloscope.

The signal output from the detector will have the same shape as that in the previous case. It will be constant in amplitude, bunched at the ends, and spread in the center. By the use of a sine-wave signal for horizontal deflection, the vertical-deflection input voltage will be traced out on the oscilloscope screen as shown in Fig. 16-33A. The bunching at the ends and spreading at the center will still exist. However, although it may not be readily evident, the variation in frequency will be *linear* this time, whereas the variation was *nonlinear* with the sawtooth sweep.

Frequency Response of Oscilloscope

With these basic factors undersood, let us now change the frequencies of both detector input signals of Fig. 16-33A. If the unmodulated signal is changed

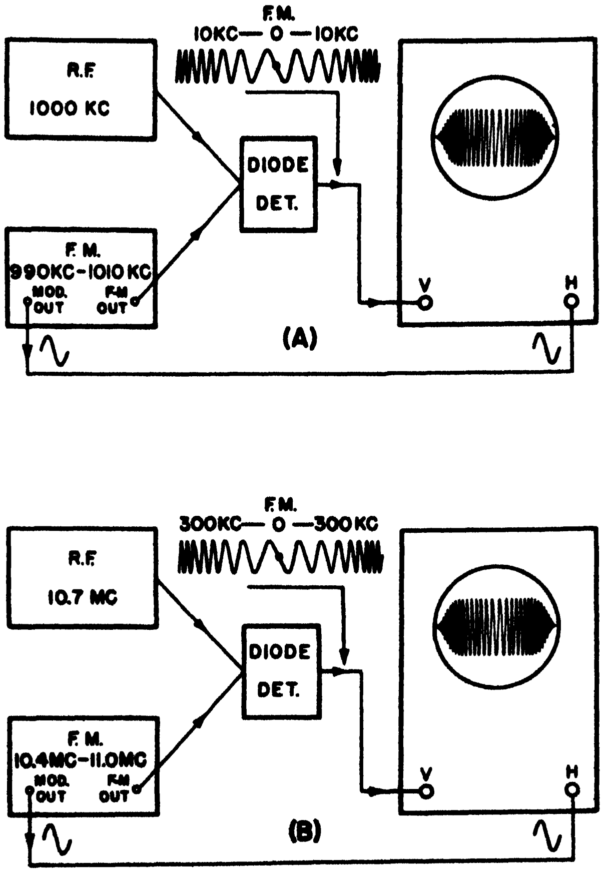


Fig. 16-33.—Here the oscilloscope is added to the circuit; its horizontal input is the modulating signal from the sweep generator. In (A) the frequencies used are the same as in Fig. 16-32; in (B) the only change is in the input frequencies.

to 10.7 Mc (which is the i.f. of many f-m receivers) and the f-m signal to 10.4 to 11.0 Mc, the resulting f-m signal output from the detector will now vary in frequency from 300 kc to 0 to 300 kc as shown in Fig. 16-33B. This new f-m signal output from the detector is generally of the same shape as that indicated in Fig. 16-33A, that is, of constant amplitude with bunching at the ends and spreading at the center, but its frequency will be different. The *frequency response* of the vertical amplifiers of the oscilloscope actually determines how much of the pattern will appear on the screen.

If the upper limit of the frequency response of these amplifiers is always higher than the highest instantaneous frequency of the input f-m signal, the complete f-m wave will always appear on the screen. Thus, for the example cited above, if the vertical amplifiers have a response of about 300 kc, the pattern that will appear on the oscilloscope is as shown in part (A) of Fig. 16-34. Points *a* indicate the maximum instantaneous frequency of the f-m beat signal (300 kc) and *b* represents the point of zero beat. For a particular setting of the horizontal-amplifier gain control, the distance between points *a-a* is due to the horizontal-deflection voltage. Since this signal is the same sine wave as the modulating signal of the sweep generator, the distance *a-a* is also said to represent the peak-to-peak frequency swing of the f-m signal, which in this particular example is 600 kc between point *b* and either point *a* the instantaneous frequency of the f-m signal increases from zero to a maximum of 300 kc.

Let us assume that we can decrease the vertical response of the oscilloscope or employ units that have a lower vertical response than that used to reproduce the pattern of Fig. 16-34A. First, assume the new response to have an upper limit of 200 kc. This means that not all of the f-m beat signal will be reproduced because its maximum frequency is greater than that of the vertical amplifier's response. The pattern that will appear on the screen will be similar to that shown in part (B) of Fig. 16-34. Since the oscilloscope will pass a maximum frequency of only 200 kc, those input signal frequencies between 200 and 300 kc will not be passed by the amplifier, and hence will not cause any vertical deflection. Consequently only the horizontal-deflection voltage is effective during the times the signal is between 200 and 300 kc and that is the reason why the pattern, between points *a-c* and *c-a*, is just a horizontal line. The remainder of the pattern is such that the frequency of the f-m wave at points *c* is about 200 kc and decreases to zero at point *b*.

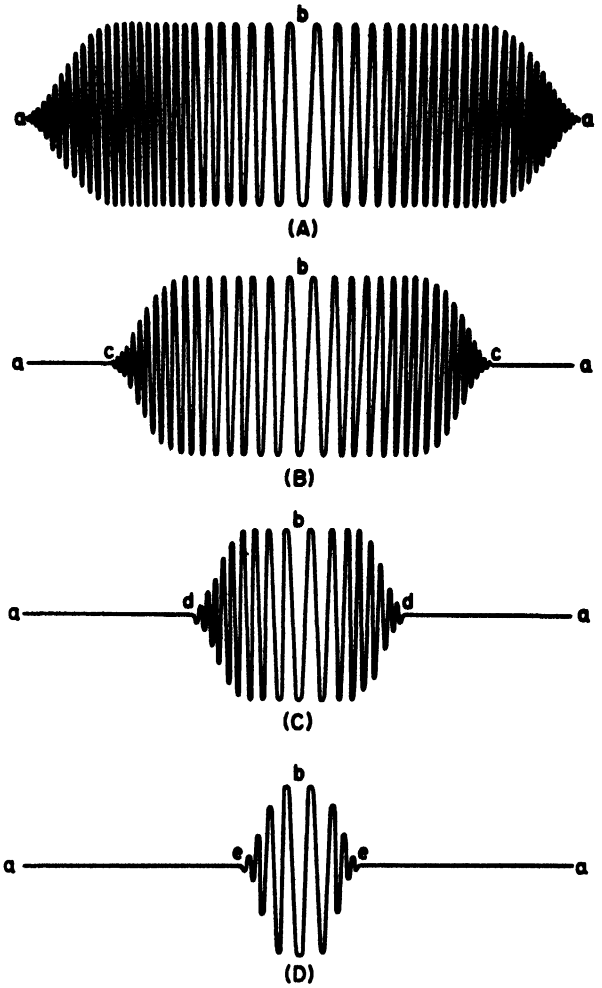


Fig. 16-34.—When the vertical amplifier response includes the highest instantaneous frequency of the f-m signal, the entire signal will appear on the oscilloscope, as shown in (A). As the response is reduced, less of the signal will appear, as shown in (B), (C), and (D), respectively.

Now let us change the vertical frequency response of the oscilloscope to a maximum of 100 kc. The pattern that results is shown in part (C) of Fig. 16-34. Less of the f-m signal is reproduced and you will note an increase in the length of the horizontal lines. Points *d* in the drawing represent the instantaneous frequency of 100 kc; frequencies above 100 kc will not be passed by the vertical amplifier. If the vertical response is reduced to a maximum of only 50 kc, then the resulting pattern appears as shown in part (D) of Fig. 16-34 where points *e* represent the 50-kc instantaneous frequency of the f-m signal. All the drawings of Fig. 16-34 are not drawn to scale and are somewhat exaggerated solely for illustrative purposes.

From the above analysis, we see that the *greater* the ratio between the maximum instantaneous frequency of the f-m input signal and the upper-frequency limit of the oscilloscope's vertical amplifiers, the smaller will be the fraction of the reproduced f-m signal. On the other hand, if the ratio is unity or less, the complete f-m beat signal will be reproduced and it will occupy the entire trace. The reproduction of the beat signals on the oscilloscope, in the manner indicated by the previous drawings, is, of course, predicated upon the assumption that there is sufficient amplification of the vertical input signal to cause appreciable deflection without overloading.

Beat Signal as a Marker

From what has been said thus far, we can see that if the beat signal is made sharper and sharper, in the manner of Fig. 16-34, it can be used as a marker to check the sweep width of the f-m signal generator. This can be done by simply varying the frequency of the generator producing the unmodulated r-f signal and noting at which frequencies the marker pip reaches the ends of the trace. These frequencies mark the upper and lower limits of the sweep width. If, for a given oscilloscope, the reproduced f-m beat signal is not sharp, then by inserting a suitable capacitor across the vertical input terminals, the signal may be made sharp enough to be used as a marker. What happens is that the capacitor bypasses to ground the high-frequency components of the beat signal. It should be remembered that these high-frequency components occupy the extreme ends of the beat pattern, and as we progress nearer the center of the signal, the frequencies of the components decrease.

Let us assume an arrangement similar to Fig. 16-33, but where the sweep generator is sweeping from 20 to 30 Mc and the unmodulated signal is 27 Mc. The resultant f-m beat note output from the detector will vary from 7 Mc to zero beat, then to 3 Mc. Without the f-m beat signal applied to the vertical terminals, a horizontal straight line will appear on the oscilloscope, the extreme ends of which represent the 20- to 30-Mc width of the sweep generator's output signal. With the f-m beat signal applied to the vertical plates, the zero-frequency beat note will appear at the 27-Mc point on the horizontal trace. How many of the other frequencies of the beat note appear depends upon the frequency range of the oscilloscope's vertical amplifiers.

In Fig. 16-35 is an analysis similar to that of Fig. 16-34 but based upon the frequencies just mentioned. (A sweep width of 20 to 30 Mc is generally employed in television receiver i-f alignment.) Part (A) illustrates the horizontal straight line that appears when only the horizontal-deflection voltage is present. The 20- and 30-Mc markings indicate the extreme frequencies of the output of the sweep generator. When an oscilloscope with a vertical-amplifier high-frequency response of 5 Mc is used and the f-m beat note is applied, the resultant pattern will appear as shown in part (B). Since the maximum response is 5 Mc, the oscilloscope cuts off at this frequency and beat-note components higher than this cannot be reproduced. When the vertical response of the oscilloscope has upper limits of 1 Mc and 100 kc, the respective patterns of parts (C) and (D) result. Note how sharp the beat note is in part (D). It appears practically as a vertical line intersecting the horizontal line. This is what is ideally desired when utilizing an unmodulated signal as a marker.

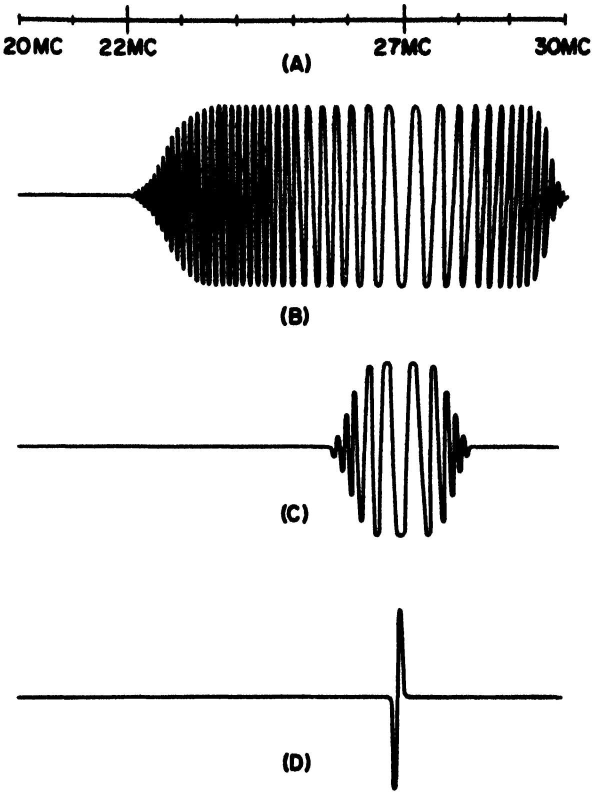


Fig. 16-35.—This is analogous to Fig. 16-34, but it is based upon television rather than f-m frequencies.

Marker As Used in Alignment

Marker signals find their greatest use in checking frequency points in wide-band resonance curves. Thus they find extensive use in television receiver alignment. They are also employed in visual alignment of f-m receivers but not to the same extent as they are in television. Marker signals are seldom used in a-m receiver alignment except perhaps in high-fidelity receivers that have double- or triple-peaked response curves.

In Fig. 16-36 are six different types of response curves with small vertical lines indicating points that may be checked by marker signals. Parts (A), (B), and (C) are symmetrical curves whereas (D), (E), and (F) are unsymmetrical. In all curves it is usually desired to check the frequency limits of the response. These are indicated by the vertical marks on the extreme ends of the curves. Aside from these points, the peaks and troughs of symmetrical and unsymmetrical response curves are also often checked.

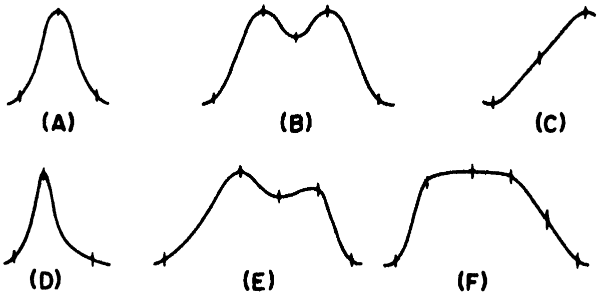


Fig. 16-36. — Six different response curves, with possible marker check points indicated.

Curve (A) is a symmetrical single-peaked response that may be found in various types of receivers. Double-peaked curves (B) and (E) are those usually encountered in overcoupled stages. Curve (C) is a typical "S" shaped characteristic as found in ratio detector or discriminator stages of f-m receivers, and f-m sound sections of television receivers. The center marker is used to identify the resonant frequency of the discriminator transformer. Curve (D) can represent a typical asymmetrical single-peaked response of an i-f transformer in stagger-tuned systems. The response of part (F) is the ideal over-all video i-f response curve of television receivers. This curve is the most important one in which marker signals are used because its frequency versus amplitude response must be fairly accurate. The end markers are not only used to identify frequency limits of the curve but also to indicate certain trap frequencies. All the other marker points are equally important. This will all be obvious in the later section of this chapter specifically dealing with television receiver visual alignment.

The block diagram of Fig. 16-37 illustrates how marker signals are used in visual alignment of tuned circuits. The only difference between this drawing and that of Fig. 16-33 is the insertion of a tuned amplifier system between the diode detector and the output of the generators. This new block may represent one or more tuned amplifier stages.

In the drawing the marker signal and f-m sweep signal are indicated as being fed to the same input point. If the marker generator and sweep generator are part of the same unit such a setup always exists normally but where the marker signal can be used independently, its exact feed-in point may be different from that indicated. Since more than one tuned stage may be used, the marker signal may be fed at any one of a number of places, besides the sweep signal input feed point. Whatever the case may be, there are two important considerations to bear in mind when using marker signals. First it is important that the coupling between the output of the marker generator and the point of its insertion in the system to be aligned be loose enough so as not to cause any undesired interaction between the marker generator and the f-m sweep generator. Second it is important not to use too much marker-signal output as this may overload the amplifiers and hence distort the alignment curve. Its output voltage should be controlled during the entire process of alignment.

When the f-m input signal sweeps over the tuned circuit, or circuits, under alignment, a response curve is traced out on the screen of the oscilloscope. This curve, as we already know, is an indication of amplitude versus frequency with frequency represented along the horizontal axis. It is often desirable to find

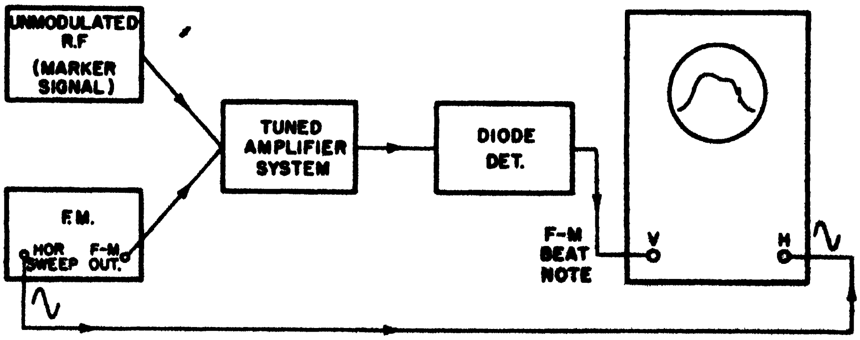


Fig. 16-37.—Block diagram illustrating the use of a marker signal in the visual alignment of tuned circuits.

out what the response is at one or more specific frequencies so as to check whether or not the tuned system is properly aligned and if it has the proper bandwidth. With the correct marker signal inserted, a marker point will appear somewhere along the curve. The marker generator, which is usually an accurately calibrated system, determines the frequency of the point on the curve where the marker pip appears. The sharper the pip the more accurately the point is determined. The curve on the oscilloscope of Fig. 16-37 illustrates an over-all video i-f response with a marker indication midway along the right slope of the curve.

VISUAL-ALIGNMENT TECHNIQUES

In order to reproduce the proper resonance curve on the oscilloscope there are a number of important things that should be known about the process of visual alignment. In this section we shall discuss the correct use of the sweep generator, oscilloscope, and marker generator to produce correct alignment patterns. Some common alignment difficulties will also be included. Before we go into the actual use of the equipment, we will start this section with some general remarks on alignment.

General Alignment Notes

Prior to the process of actual alignment, there are certain receiver adjustments that should be made in order to insure correct alignment. These prealignment adjustments usually mean special settings of the different controls on the receiver. In the discussion that follows, on actual a-m, f-m, and television receiver alignment procedures, the required prealignment adjustments will be mentioned. These adjustments will not always be the same for each type of receiver because of varying design features. Whatever the case may be, it is always advisable to follow the individual manufacturer's preadjustment settings, if any are given.

When manufacturer's visual-alignment instructions are supplied, these instructions follow a specific sequence. It is our suggestion that this sequence be followed whenever possible because the manufacturer knows his receiver best. However, in the absence of such data it is suggested that the following visual-alignment sequence be followed for each particular type of receiver. In all of the following cases, when alignment of the i-f sections is mentioned, the i-f stages should be aligned starting with the last stage first and working backward toward the first stage.

1. For *a-m receivers* the procedure is simple; the i-f stages are aligned first, the oscillator section next, and finally the r-f section.
2. In *f-m receivers* the detector transformer should be aligned first. The limiter stages should be aligned next, starting with the last limiter stage, if more than one limiter is used. The i-f system is then to be aligned, followed by the oscillator and r-f sections of the receiver.
3. For television receivers a different sequence of operations is used because of the numerous tuned circuits. The f-m sound section is aligned first; the detector, limiter, and i-f stages are aligned in that order. Trap circuits, if any are employed, are aligned next. The remainder of the alignment sequence is as follows: video i-f, oscillator, and finally the r-f stages of the receiver.

Care should always be taken that the sweep generator, and marker generator if one is used, be loosely coupled to the input circuit of the receiver under test. This is necessary in order to prevent possible overloading, regeneration, or oscillation.

The response curve as seen on the oscilloscope may appear upright or inverted. The polarity of the pattern does not alter the presentation of facts; either pattern is equally useful. However, for the sake of illustration, wherever possible in the remainder of the chapter, alignment patterns will be shown where the curve appears upright.

It is always advisable to have a set of insulated alignment tools; if possible all tools should be made *completely* of some plastic or fiber material, although there are some cases where metallic screwdrivers may be used for aligning certain r-f and i-f circuits. Since the size of the holes or the space available through which the adjustment is to be made may vary considerably, it is suggested that the serviceman's set of alignment tools be of various sizes both in length and diameter.

Proper Use of the Sweep Generator and Oscilloscope

In order to understand some of the techniques involved in the use of the sweep generator in conjunction with the oscilloscope, we have made a simple drawing of a typical hookup, in Fig. 16-38. This circuit may be representative of part of the i-f system of any type of receiver. In this circuit, tubes *V1* and *V2* are amplifiers, tube *V3* is the detector or some rectifying system within the receiver, and *T1* and *T2* are i-f transformers resonant to the same frequency. An f-m signal generator is schematized in the drawing. Only those parts of the circuit that are considered important are illustrated. No controls are shown on the sweep generator because of the wide variety of names and positions of these controls. Only the two essential terminal connections are shown. Common ground connections between the sweep generator, receiver, and oscilloscope are made.

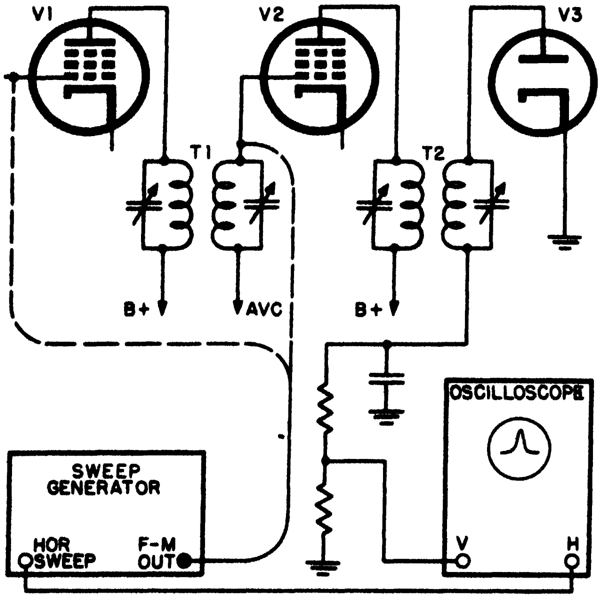


Fig. 16-38.—Block diagram illustrating the use of the sweep generator in conjunction with the oscilloscope for visual alignment.

Let us assume that each i-f transformer has a single-peaked resonance curve and that it is desired to align transformer *T2* first. The f-m signal output terminals of the sweep generator should be loosely coupled to the grid input of amplifier *V2*. If it were fed to *V1* the resonance curve appearing on the screen of the oscilloscope would be the combination of the resonance curves of *T1* and *T2* and not that of *T2* alone. The vertical terminal of the oscilloscope is connected across part of the detector load.

The modulating (sine-wave) signal output from the sweep generator (which we have labeled HOR. SWEEP) should be connected to the high side of the horizontal input terminals of the oscilloscope. The voltage output control of the f-m sweep is adjusted to the minimum voltage that will produce a satisfactory pattern, without any attenuation in the oscilloscope. This minimum f-m signal voltage output is to insure against overloading of the i-f stages being aligned. The gain controls of the horizontal and vertical amplifiers of the oscilloscope are adjusted to produce a response curve that will satisfactorily fill the screen.

The sweep-width control should be adjusted so that the *complete* resonance curve appears on the oscilloscope. In visual alignment the amount of deviation of the f-m signal output from the generator determines how much of the resonance curve will appear on the screen of the oscilloscope. It is important that the amount of deviation used is always greater than the bandwidth of the tuned circuit. This is necessary in order to see the complete resonance curve, including the parts beyond the skirts of the curve. Some sweep generators are unable to produce sufficient deviation to make visible very much of certain response curves. This need not prevent alignment if one can obtain all the necessary information from that part of the curve which does appear.

In order to see how the amount of frequency deviation affects the reproduced response curve let us examine Fig. 16-39. The four oscillograms that appear in this figure depict the response of a typical single-peaked wide-band i-f system of an f-m receiver under different settings of the sweep generator's sweep-width control. The dots appearing on the curves were inserted for the sake of illustration. For part (A) the peak-to-peak deviation employed was 600 kc, more than sufficient to see the complete response curve, including the parts beyond the skirts or sides of the curve. In part (B) a total deviation of 450 kc was used, which was still sufficient to obtain a complete curve. In parts (C) and (D) the total deviation used was 300 kc and 200 kc respectively. From these latter two curves we see that if the amount of deviation used is not sufficient, only part of the resonance curve will appear on the oscilloscope.

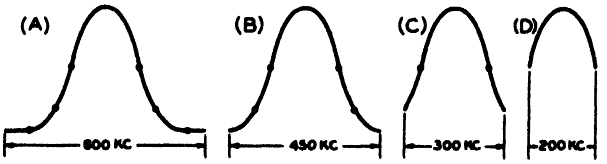
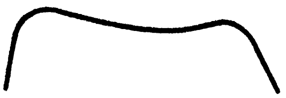
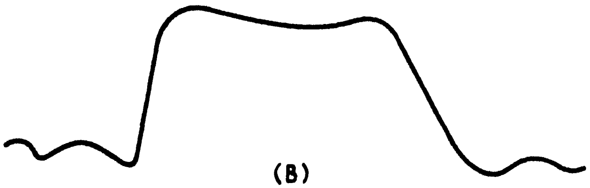


Fig. 16-39.—Illustrating the effect of the amount of frequency deviation on the appearance of the reproduced response curve.

In Fig. 16-40 the two patterns illustrate another case of insufficient sweep width. The patterns represent an over-all video i-f response. The pattern at (A) was produced when the sweep width was not sufficient. Note how important parts of the curve are omitted, when (A) is compared with (B). Once the desired amount of deviation is set by the sweep-width control on the sweep generator, the tuned circuit should then be ready for alignment.



(A)
INSUFFICIENT SWEEP WIDTH



(B)
SUFFICIENT SWEEP WIDTH

Fig. 16-40.—Another example of the effect of deviation on a response curve.

The center frequency setting is quite important, especially if the sweep width used is just about equal to the bandwidth of the tuned system under alignment. If the center frequency adjustment is too far off, either side of the response curve may not be reproduced. Refer to Fig. 16-41. The four patterns in this drawing represent those of an over-all video i-f response. The sweep-width control of the sweep generator is adjusted to approximately 10 Mc as is indicated by the two dashed vertical lines on the side of each drawing. The dotted line between the two dashed lines indicates the center of the response curve. Cases (A) and (B) indicate two conditions where the center frequency adjustment control, although slightly off the center frequency of the response, still enables the complete curve to be seen. In parts (C) and (D) a different situation exists. The frequency adjustment was too far off and only part of the curve appears. In (C) the center frequency adjustment was too high, and part of the right side of the curve was not reproduced; in (D) the center frequency was too low and the left side of the curve was not reproduced.

If the resulting pattern appears as a double image the phasing control on the sweep generator should be adjusted as best possible to produce a single-line response trace.

Once transformer *T2* of Fig. 16-38 is aligned, the f-m output cable of the sweep generator should be connected to the grid of *V1* for alignment of transformer *T1*. Now, the combined response of *T1* and *T2* appears on the oscilloscope. What was said before about manipulation of the controls on the sweep generator applies here too. Transformer *T2* is then adjusted to produce the desired response curve on the oscilloscope. If the response of *T1* alone is desired, a probe detector can be used. This means that the vertical terminals of the oscilloscope have to be disconnected from the detector of the receiver and the proper end of the probe connected to these terminals. The ground side of the other end of the probe is connected to any ground of the circuit and the high side is connected to either the grid or plate of *V2*. In this manner only the response of *T1* will appear upon the oscilloscope. If the probe is connected to the grid of *V2* care should be taken that it does not appreciably load the secondary of *T1*, as this will cause a false response curve to appear.

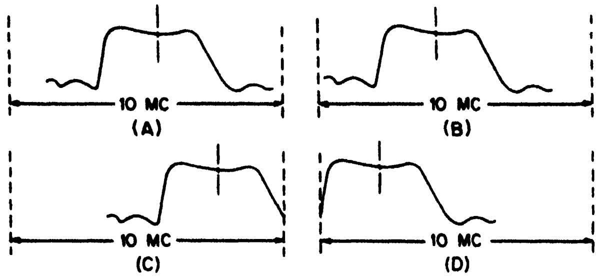


Fig. 16-41.—How the setting of the center frequency control of the sweep generator may affect the appearance of the reproduced response curve.

Varying the Marker Output

When a marker generator is used in which the output signal is obtained from a cable other than that of the f-m sweep output signal, this cable should be *loosely* coupled to the desired part of the circuit. When this marker signal is to be injected into the same part of the circuit as the f-m sweep signal both signals should be loosely coupled to the circuit in question. If the marker signal and f-m sweep signal appear at the same terminal in a combined type of generator there is no problem of marker signal insertion; it is automatically inserted at the same phase as the f-m signal.

Once the choice of the marker frequency is made, the marker amplitude control should be properly adjusted to produce the correct marker size. We previously mentioned that the output of the marker generator should not be too great as this may overload the tuned amplifier being aligned. At the same time the output of the generator should not be too small as this may not produce any noticeable marker pip. Control of the marker signal amplitude should be continuously maintained when different frequency points of an alignment curve are to be checked. This is especially important where the response of the curve changes appreciably for these different frequency points.

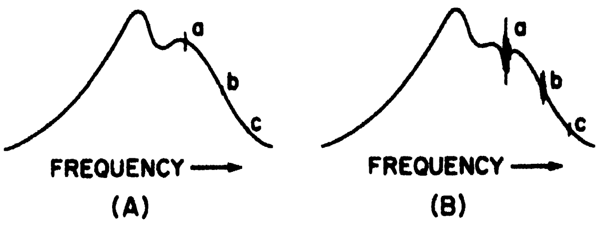


Fig. 16-42.—Illustrating the need for control of the marker amplitude. For a given marker signal level, the appearance of the marker on the oscilloscope may vary considerably, depending upon its location on the response curve.

For example, in a curve shaped like that of Fig. 16-42A, assume that the responses at points *a, b,* and *c* are to be checked, to determine at what frequencies they occur. Assume that the first point checked is *a* and that the output of the marker generator is adjusted so that it will give us a fair marker indication, as noted on the drawing at this point. If points *b* and *c* are to be checked in succession and if the output of the marker generator is held constant (the same as for marker *a*) the size of the marker indications will be reduced, with that at *c* smaller than that at *b*. This is so because the gain or response of the tuned amplifiers at these frequencies will be less than at *a*. You will note that the marker at *c* is practically nonexistent. From this analysis we see that for the case where we are checking frequency points and proceeding from a greater response level to a smaller one the output of the marker generator should be increased to allow for reasonable marker indications.

Suppose we are checking frequencies from a point of smaller response to one of greater response. The drawing of Fig. 16-42B, which is the same curve as that of part (A), also indicates the same three marker points as (A), point *c* is to be marked first. In this case the amplitude of the marker at *c* is adjusted for what is considered a satisfactory indication. If the amplitude of the marker output remains approximately the same but its frequency is reduced to points *b* and *a*, the marker amplitude that appears will be increased in each case because of the increase in response at these frequencies. In the case under discussion the marker signal for the indication at *a* may be great enough to cause overload of the amplifiers, and hence distort the curve as indicated in the drawing of part (B).

Both conditions of Fig. 16-42 tell us that the output of the marker generator should be continuously controlled to produce a marker just sufficient to give a readable indication. The only method of avoiding such continuous control of the marker-generator output is to have a system whereby the marker signal is not passed through the tuned amplifiers being aligned. Thus it will not be subject to the response of the tuned circuits. Such systems have been designed; one of these employs a special marker generator (see Figs. 14-34 and 14-35 in Chapter 14), which can be used with any sweep generator. Only the f-m sweep signal is fed to the amplifiers being aligned. This f-m sweep signal is also fed to a special input terminal on the marker generator. The sweep signal and marker signal beat together inside the marker generator unit to produce the difference beat frequency.

The signal output from the detector circuit of the receiver being aligned is also fed into the marker-generator unit but at a different point. This latter signal, which represents the amplitude versus frequency response of the tuned system under test, is combined with the f-m beat signal inside the marker generator and then fed to the vertical amplifier of the oscilloscope.

Whatever type of marker generator is to be used in the process of visual alignment, it should meet certain standards to be truly useful. As mentioned, its frequency calibration must be fairly accurate and should remain quite stable after an initial warm-up period. The output should not be subject to variation in frequency when the load on the generator is changed. The voltage output of the generator should be susceptible of approximately the same degree of variation for most frequencies.

One of the most important requirements of a marker generator is that it should be capable of covering a suitable range of frequencies for the particular job for which it is to be used. Its most important frequency range is that of the i-f circuits in receivers; r-f ranges are generally secondary in importance although many generators are so designed that they are capable of both ranges.

Some Common Alignment Difficulties

The process of visual alignment has its share of difficulties as do other methods of servicing. In this section we will study some resonance curves illustrating a few of the common faults encountered. The majority of pictures included in this section were taken from an a-m superheterodyne receiver but they can, nevertheless, be used as a general indication for all types of receivers.

Insufficient Coupling Capacitance

Certain forms of circuit connections between the cathode-ray oscilloscope and the source of the rectified signal required for alignment introduce phase distortion of the audio signal, with the result that the resonance curve is distorted. Such distortion is very often introduced when the rectified signal from the receiver is fed to the vertical terminals of the oscilloscope through too small a coupling capacitor. This is usually the case when the vertical-deflection voltage is secured from the audio volume control and said control is connected to the source of the audio signal through a small blocking capacitor.

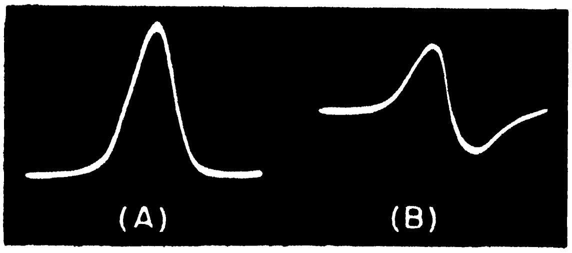


Fig 16-43.—(A) is a single-peaked normal response curve; (B) shows phase distortion of the same response due to insufficient coupling capacitance.

Due to this small capacitance, the resulting alignment pattern, instead of having its base upon one horizontal line has one side of the base correctly placed and the other side displaced below this line. This is shown in Fig. 16-43. Part (A) of this figure illustrates a normal single-peaked resonance curve whereas part (B) shows the resonance curve for the same circuit but with distortion. In the double-peaked patterns of Fig. 16-44 part (A) represents the normal curve whereas part (B) indicates phase distortion due to insufficient coupling capacitance.

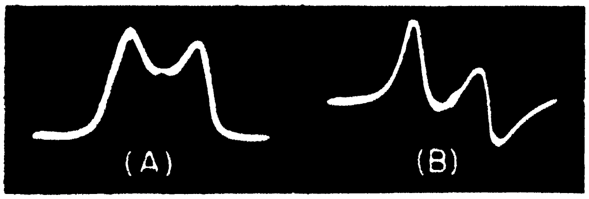


Fig. 16-44.—Same as Fig. 16-43 except for a double-peaked curve.

Excessive F-M Signal

In the description of the process of visual alignment, it was suggested that the strength of the f-m signal output from the sweep generator be made just great enough to produce a satisfactory pattern size on the oscilloscope. If the input f-m signal is too strong it may overload the receiver tubes and cause erroneous alignment patterns. The photographs of Fig. 16-45 illustrate this. In part (A) is a single-peaked resonance curve illustrating a proper amount of f-m signal input. When the f-m signal input is increased to the point of overloading the tubes, the alignment pattern changes to that illustrated in part (B). A further increase in f-m signal causes a greater degree of overload as shown by the oscillogram at (C).

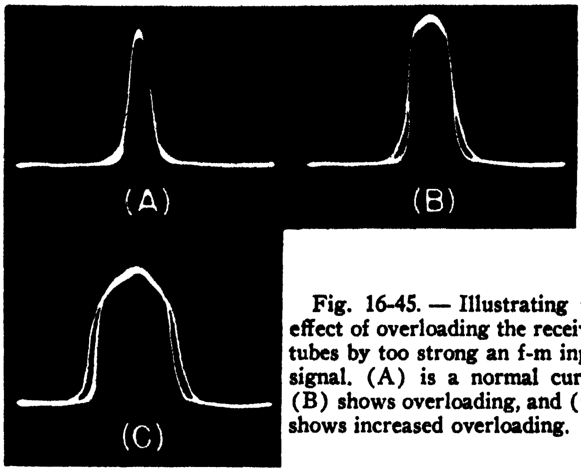


Fig. 16-45. — Illustrating the effect of overloading the receiver tubes by too strong an f-m input signal. (A) is a normal curve, (B) shows overloading, and (C) shows increased overloading.

Comparison of these three curves indicates that, for overload conditions, the resonance curve becomes broader, giving a flattened wideband appearance. This occurs because the overloading causes grid current to flow; this effectively makes the resonant circuits broader and also may cause a change in the resonant frequency of the tuned circuits. It is interesting to note the relative amplitudes of these three images, bearing in mind that the automatic-volume-control signal was present in the receiver at all times.

Regeneration

In some alignment situations the input signal may not overload the tube, but due to some circuit coupling,

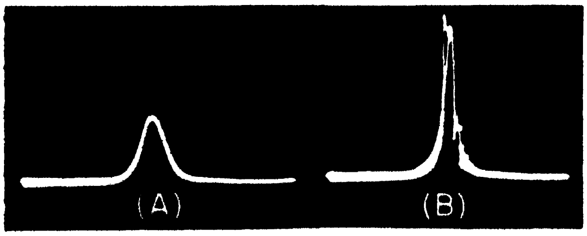


Fig. 16-46.—(A) is a normal response curve; (B) is the same curve with regeneration occurring.

regeneration may occur which may cause false alignment patterns. For example in Fig. 16-46 oscillogram (A) indicates a correct alignment curve with no regeneration present and pattern (B) occurs when a certain amount of regeneration is present. The regeneration in this case has distorted the right slope of the resonance curve, increased its amplitude, and made the response less broad.

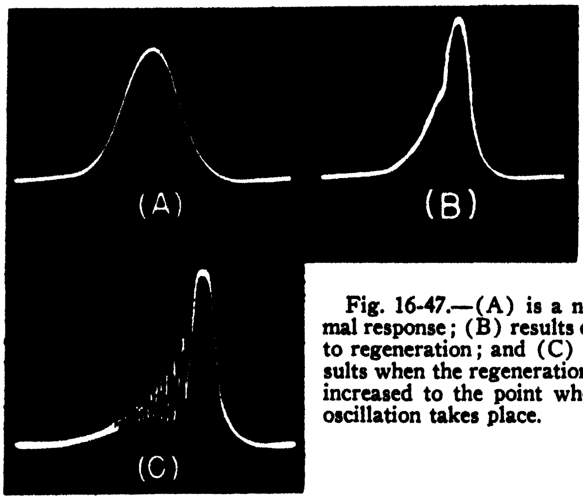


Fig. 16-47.—(A) is a normal response; (B) results due to regeneration; and (C) results when the regeneration is increased to the point where oscillation takes place.

The effects of regeneration are also indicated by the patterns of Fig. 16-47. The photograph at (A) indicates the correct single-peaked symmetrical resonance curve. Only a minimum amount of regeneration was present when this curve appeared. With increased regeneration the resonance curve took on the shape of pattern (B). The increased regeneration increased the amplitude of the curve, made the response sharper, and caused the over-all shape to become asymmetrical. Note the nick in the left slope of this curve. When the regeneration was increased to the point where oscillations occurred, the resonance curve of part (C) results. The oscillations are evident in this photograph.

Spurious Voltage

In the course of alignment, undesired voltages may be picked up and may interfere with the resonance curve produced on the oscilloscope. Such spurious voltages can be very detrimental to the interpretation of the alignment curve. In this section we shall study some resonance curves containing undesired signals.

If there is insufficient r-f filtering action in the detector circuit of the receiver being tested, some of the carrier voltage may be picked up and fed to the vertical-deflection plates of the oscilloscope along with the voltage representing the response curve. A typical pattern with such interference is shown in Fig. 16-48. This pattern was produced when the bypass capacitors in the filter network of the diode detector were removed. The presence of the carrier voltage in this instance produces a fuzzy effect, broadening the lines of the curve. It should be remembered at this point that the r.f. can appear only if the vertical amplifiers of the oscilloscope have the proper frequency response.

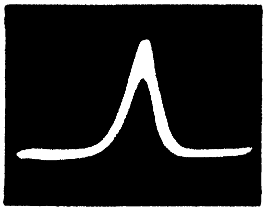


Fig. 16-48.—Response curve containing r.f., due to insufficient filtering.

The distorted alignment pattern of Fig. 16-49 is due to the presence of an a-m wave modulated at 400 cps being fed through the i-f amplifiers. This picture occurred when a 400-cps a-m switch, part of the sweep generator used, was turned on in addition to the f-m switch. Note how this interference prevents correct interpretation of the alignment pattern. We cannot tell whether the circuit under test has a single- or a double-peaked response curve or if it is symmetrical or asymmetrical.

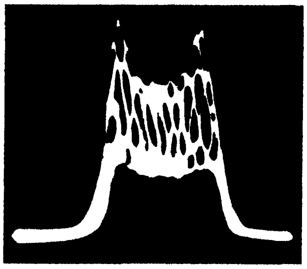


Fig. 16-49.—Response curve with distortion due to the presence of a 400-cps a-m wave in the i-f amplifiers.

If an r-f voltage within the frequency range of the tuned circuit being aligned finds its way into the tuned circuit, it will beat against (i.e. heterodyne) the f-m sweep signal in the detector of the receiver, and interfere with the resultant pattern (an action similar to that which occurs for marker indication). The single-peaked resonance curve of Fig. 16-50 illustrates such a

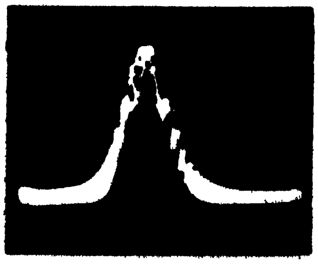


Fig. 16-50.—Distortion due to a spurious r-f voltage, equal to the i.f., beating against the f-m wave in the detector.

condition. The frequency of the interfering r-f signal in this case was at the resonant frequency of the tuned circuit being tested. This signal, in beating against the f-m signal, will produce a varying beat note that is the difference frequency between the two signals. A zero beat note will occur at the center frequency of the tuned circuit whereas at other frequencies of the f-m signal, the beat note will be different from zero.

Let us examine this situation a step further. Assume that the resonant frequency of the tuned circuit and also that of the interfering signal is 455 kc and that the input f-m sweep signal is varying between 450 and 460 kc. This means that the resultant difference frequency beat note output from the detector of the receiver will be varying in frequency from 5 kc to 0 and then to 5 kc. If the vertical amplifiers of the oscilloscope have a frequency response capable of passing these difference frequencies, then they will appear superimposed upon the alignment curve with the higher frequency components at the far ends of the curve and the lower ones toward the center, with the zero beat note occurring at the center. This is what has happened in the curve of Fig. 16-50. The zero beat note is not very clear from this photograph but careful examination will reveal it.

Let us assume another interfering unmodulated r-f signal, this time at a frequency several kilocycles away from the center frequency of the f-m signal, yet within the same 450 to 460 kc pass-band of the tuned circuit in question. This interfering signal will also beat against the f-m signal in the detector but the zero beat note this time will occur somewhere along the side of the single-peaked curve. If a sweep generator employing sine-wave sweep is used then the zero beat note will occur on only one side of the curve. On either side of this zero beat note, other beat components will exist which increase in frequency as we move further away from the zero beat note. For example let us assume that the interfering r-f signal is 458 kc, that the f-m signal sweeps 450 to 460 kc, and that the resonant frequency of the tuned circuit is 455 kc. This means that the resultant f-m beat note will vary from 8 kc at the low end of the curve (the 450-kc point) to 2 kc at the high end of the curve (the 460-kc point), passing through a zero-beat point at the 458-kc point of the curve. At the resonant frequency of the tuned circuit the beat component will be 3 kc.

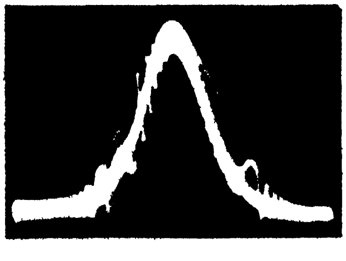
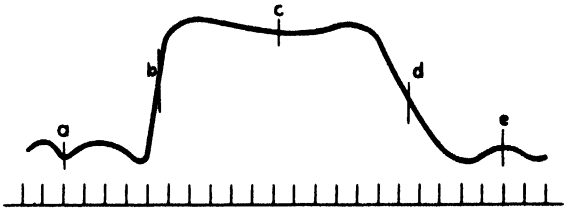


Fig. 16-51.—Similar distortion as in Fig. 16-50 but due to the frequency of the interfering signal and because a wobbulator is used, a double beat note appears, one on each side of the curve.

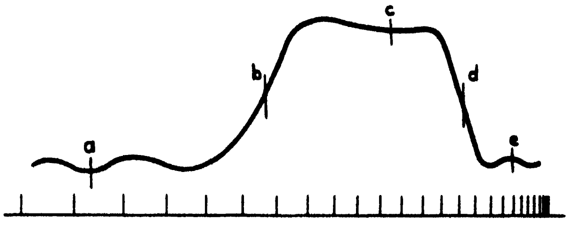
If a wobbulator type of sweep generator is used, the zero beat will appear on both slopes of the single-peaked curve, because of the manner in which the wobbulator sweep signal varies in frequency. A photograph of a single-peaked curve illustrating this condition of interference appears in Fig. 16-51. Note the zero beats on both sides of the curve.

Nonlinear Sawtooth

There are some sweep generators that utilize a sawtooth signal as the modulating voltage of the f-m signal. This means that with such units a phasing control is not necessary and the sawtooth time base of the oscilloscope can be used as the horizontal-deflection voltage. If either of these sawtooth voltages is distorted, and hence nonlinear, a defective alignment pattern will result.



(A) NORMAL PICTURE
(LINEAR SWEEP)



(B) DISTORTED PICTURE
(NON-LINEAR SWEEP)

Fig. 16-52.—(A) illustrates a normal over-all video i-f response curve; (B) is the result of what happens due to a nonlinear sawtooth sweep.

Such distortion is indicated in Fig. 16-52. Part (A) indicates a normal video i-f alignment curve. The graduated frequency marks below the curve indicate the linearity of both sawtooth signals. The pattern of part (B) resulted when one of the sawtooth signals became nonlinear. The frequency marks at the bottom indicate that the pattern is spread out on the left and bunched on the right. Letters *a* to *e* are markers that indicate corresponding frequency points in each curve.

VISUAL ALIGNMENT OF A-M RECEIVERS

We are now ready to study the methods of visual alignment of a-m, f-m, and television receivers. While visual alignment of a-m receivers is not practiced to the same extent as it is in f-m and television receivers, it is important enough, especially where wide-band circuits are encountered, to be included in this section. Alignment of the r-f and i-f stages of all three types of receivers will be discussed.

Visual alignment of a-m receivers will be considered first. Many of the general comments on alignment procedures made in this section will also hold true for visual alignment of f-m and television receivers. For example, in each case the process of alignment will start at the detector and we will work backward toward the i-f and finally the r-f sections of the receiver. The types of detector circuits found in the receivers will be discussed first because the frequency versus amplitude characteristic (i.e. the response of the tuned circuits) appears in the output of the detector circuit and it is this circuit to which the vertical terminals of the oscilloscope are to be connected.

Detector Circuits

Most a-m receivers today employ diodes as detectors. Triode and pentode detector circuits are also found but only to a limited degree. However, in order to be complete, the two latter types will be briefly included in our discussion. In Fig. 16-53 four such circuits are illustrated. Circuits (A) and (B) use triodes as detectors and circuits (C) and (D) employ pentodes. Circuits (A) and (C) use cathode bias and (B) and (D) use grid-leak bias. In all circuits resistor *R1* represents the load for each tube and capacitor *C1* is employed to bypass the high frequencies to ground. Point *V* in each drawing indicates the point to which the high side of the oscilloscope's vertical input terminals is to be connected. This point is chosen because, during visual alignment, the rectified and filtered output signal, which is varying in amplitude with respect to time, appears across the load resistor *R1*. This signal output represents the frequency response of all the tuned circuits through which the f-m sweep signal passes. The additional capacitor *C3* and inductor *L* in Fig. 16-53B together with capacitor *C1* form a pi-section r-f filter. Such filters as this are not found to any great extent today; it is merely included to indicate a type that does exist.

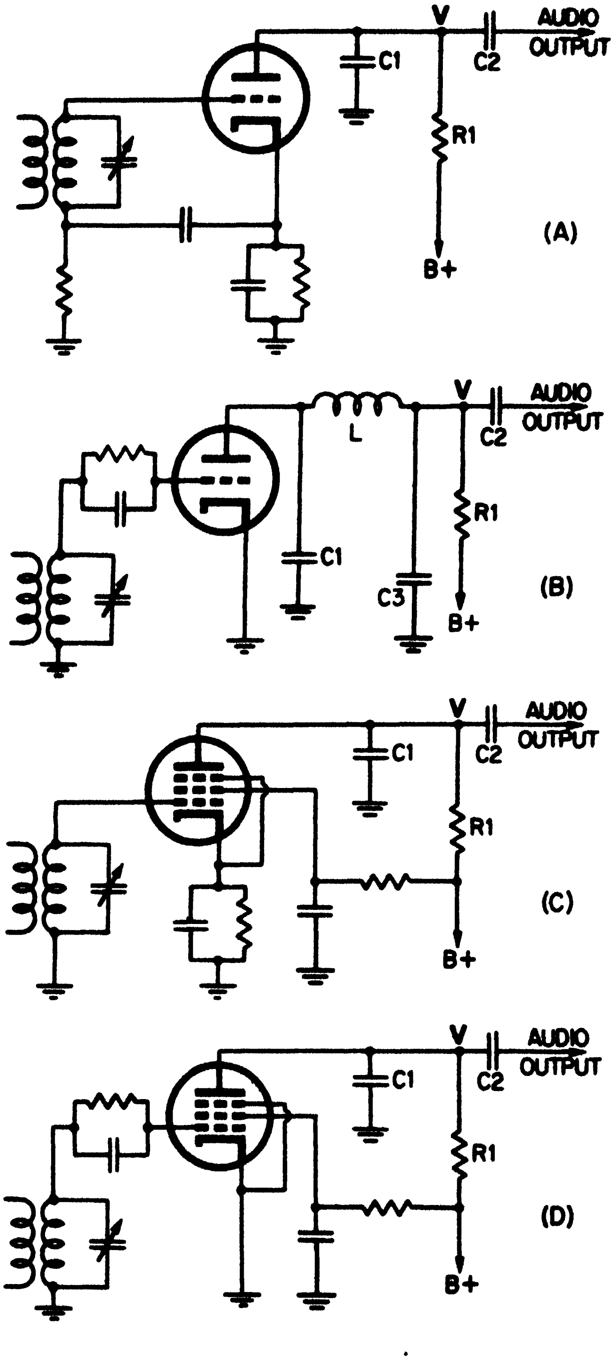


Fig. 16-53.—Typical a-m detector circuits using triodes and pentodes.

Capacitor *C2* in each circuit is used to couple the audio signal to the following stages. The vertical terminals of the oscilloscope *should not* be connected at the output of this capacitor as this may distort the alignment pattern if the capacitance is too small. Detector circuits such as those indicated in Fig. 16-53 can be used in both trf and superheterodyne receivers. This means that the tuned circuits at the input to the detector will be resonant to r-f signals when used in trf receivers and to i-f signals when used in superheterodyne receivers.

Let us now examine a typical diode detector circuit as used in practically all of the superheterodyne a-m receivers manufactured today. The diagram appears in Fig. 16-54. Although the diode is illustrated as a separate tube it may be part of a multipurpose tube, such as a diode-triode or a duo-diode. In fact the majority of a-m receivers today employ diode-triode or diode-pentode tubes (or duo-diode triode and duo-diode pentodes) where the diode section is used for detection and the triode or pentode section is used as the first audio amplifier. Since we are only interested in the diode detector circuit it is shown as existing alone inside the single envelope, for simplicity.

The double-tuned circuit in the drawing represents the last i-f transformer of the receiver. The *R1-C1-C2* network is a simple circuit to filter out the i-f component of the rectified signal. The output of this network is the audio signal. Components *R2* and *C3* constitute an audio filter and the output of this circuit is used as an avc voltage. The potentiometer functions as the volume control of the receiver and *C4* is used to couple the audio signal output to the following audio amplifier.

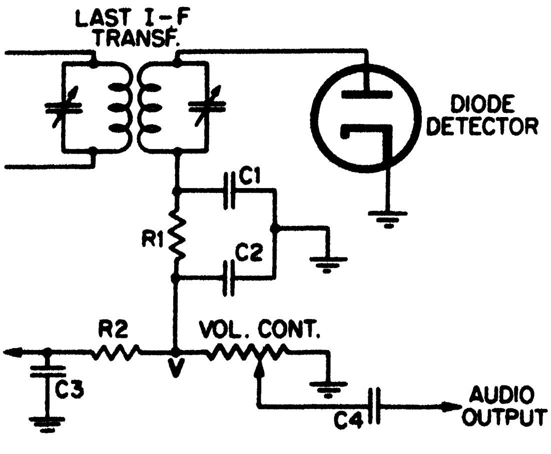


Fig. 16-54.—Typical a-m diode detector circuit.

Point *V* indicates the place where the high side of the vertical input terminals of the oscilloscope is to be connected for visual alignment. The rectified output signal appears at this point. If the oscilloscope is placed at the junction of *R1* and *C1* the resultant alignment pattern may appear fuzzy due to the presence of some i-f carrier voltage. If the oscilloscope is placed to the right of capacitor *C4* the pattern may become distorted because the value of *C4* may not be large enough.

I-F Alignment

Since most manufacturers list the points of the i-f stages that they desire the f-m sweep signal to be fed into (which may be closely coupled or directly connected to the grid of the amplifier preceding the i-f transformer to be aligned) we will not schematically show such points in our discussion. A block diagram of a typical a-m receiver will be used instead of a complete schematic circuit. In the visual alignment of the i-f stages, the oscillator tuning control can be set at the high end of the band in order to prevent any interference. If oscillator interference exists then it would be wise to make the oscillator inoperative by simply grounding its grid circuit. This can be done by simply placing a jumper between the grid and ground.

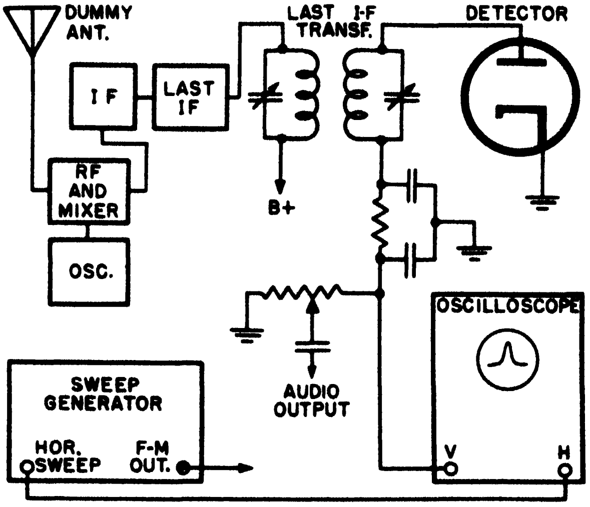


Fig. 16-55.—Block diagram of basic setup for visual alignment of an a-m receiver.

The block diagram of Fig. 16-55 illustrates a basic setup for visual alignment of an a-m receiver. Only the detector circuit and the last i-f transformer are shown in schematic form. The oscilloscope's vertical input (marked *V* in the drawing) is connected to the detector load and remains there ***throughout*** the visual-alignment process of the receiver. The sine-wave sweep of the generator (from terminal marked HOR. SWEEP in the drawing) is connected to the horizontal input of the oscilloscope (marked *H* in the drawing). The internal sweep of the oscilloscope is not used. Common ground connections between the generator, receiver, and oscilloscope should be made.

The response of the i-f stages in the majority of a-m receivers is a single-peaked curve. Only in special cases, as in high-fidelity receivers, will it be double-peaked, triple-peaked, or some other shape. Double-tuned transformers are used in the i-f stages of most present-day a-m receivers. In aligning these i-f transformers the secondary is generally tuned first and the primary last. This is the procedure which will be referred to here.

The last i-f stage of the receiver is aligned first. The f-m output cable of the sweep generator is connected to the grid of the last i-f amplifier. The frequency control of the sweep generator should be set to the i.f. of this transformer. The sweep width of the f-m signal should be sufficient to include the complete curve, but not too great or the curve may appear cramped upon the screen and thus the pattern will be difficult to interpret.

If the i-f transformers all have single-peaked resonance curves, it is a simple matter to align them. In each case the adjustment is made so that a single-peaked resonance curve (usually symmetrical) of maximum amplitude appears on the oscilloscope. If the manufacturer lists the response of the transformer as being different from a single-peaked curve then the transformer should be aligned to produce the shape of curve suggested.

All the i-f transformers of the receiver are aligned in similar fashion. The only change in the circuit hookup is that in each case the sweep generator cable is coupled to the grid of the i-f amplifier preceding the stage previously aligned. One thing should be borne in mind during the alignment procedure as one works backward from the last i-f stage to the first; that is, the amplitude control of the f-m signal output should always be regulated to prevent overloading of the i-f stages. This is necessary because an extra stage of i-f amplification is in the signal path every time a new i-f stage is aligned.

If one of the i-f stages has a resonance curve that is not single-peaked, then the resulting response curve on the oscilloscope will have some other shape. The exact shape should be known so that proper alignment can be made. This is especially true in high-fidelity receivers. In Fig. 16-56 appears a triple-peaked symmetrical over-all response curve of the i-f section of a high-fidelity a-m receiver. This response curve is the result of the combination of a single-peaked and a double-peaked resonance curve where each has the same center frequency. What happened is that the single-peaked resonance curve, when added to the double-peaked resonance curve, raised the response at the center frequency, causing a "lifting" of the trough of the double-peaked curve to produce the small central peak shown in the photograph.

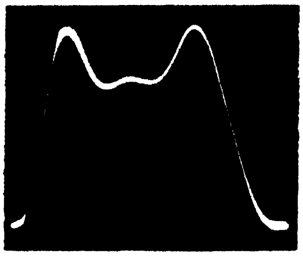


Fig. 16-56.—Triple-peaked symmetrical over-all i-f response curve of a high-fidelity a-m receiver.

Examination of Fig. 16-57 indicates three different types of over-all i-f response curves of a-m receivers. (A) is a symmetrical single-peaked curve which is obtained from the usual type of a-m broadcast receiver whereas (B) and (C) are obtained from broadband i-f circuits of high-fidelity receivers. Both of these latter curves are unsymmetrical. Curve (C) contains a double peak. The three curves illustrated in Fig. 16-57 are those requested by the designers of three different types of receivers.

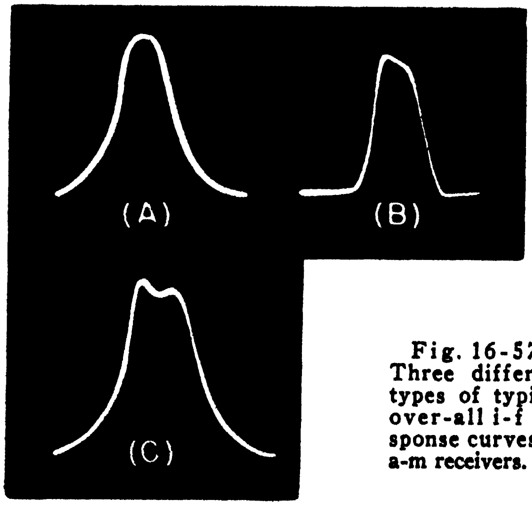


Fig. 16-57.—Three different types of typical over-all i-f response curves of a-m receivers.

An alternate method of i-f alignment, especially of a-m receivers where each i-f stage has the same response, is to keep the sweep generator cable at the grid input to the converter tube (or mixer tube) throughout the i-f alignment. The frequency of the generator remains set at the i.f. of the receiver and each i-f transformer is aligned starting from the last stage and working backward. Each stage is aligned for the proper type of response curve. During this process, the amplitude control of the sweep generator may require regulation. This method should not be used if the i-f stages are completely out of alignment but only if they are slightly so. Whatever method is used it is usually suggested that the process be repeated for a more accurate alignment.

Oscillator and R-F Alignment

It is a simple matter to visually align the oscillator and r-f stages once the i-f stages are aligned. The oscillator system is usually aligned before the r-f stages. The oscilloscope connections remain as they were for the i-f alignment, as shown in Fig. 16-55, but the sweep generator's f-m output cable connection is changed. The cable is coupled to the dummy antenna as suggested by the manufacturer.

The alignment of the r-f and oscillator sections differs somewhat with different manufacturers. Some suggest starting at the high end of the band and working down, while others stipulate that alignment should start at the low end of the band; still others suggest going from end to end and finally to some middle point of the band. The result will be essentially the same if the complete alignment instructions are followed. It should be remembered that some circuits have more trimmers and tuning slugs than others. By the use of these extra adjustments, a finer alignment can be obtained over the entire band as long as we know which variable components to tune. For purposes of discussion we will start at the high end of the broadcast band.

If the oscillator had previously been made inoperative, this stage should now be put back into operation. The frequency control of the sweep generator is set to a center frequency of about 1,500 kc, and the receiver dial is also set to this frequency. The sweep-width control of the sweep generator is so adjusted that a little more than the complete response curve appears on the oscilloscope. The necessary *oscillator* trimmers and/or tuning slugs are then adjusted to produce a maximum response. Remember that the response curve seen during this stage of the alignment is a combination of the over-all i-f response curve plus the response of the r-f stages (which usually have a greater bandwidth than the i-f stages), and hence may not have the same shape as that of the i-f system alone.

After the oscillator is adjusted, the necessary *r-f* trimmers and/or tuning slugs are varied for maximum response on the oscilloscope. If there is more than one r-f tuned circuit, adjust the one used as the input to the mixer or converter first and then adjust the others, working backward to the antenna. Usually not more than two r-f tuned circuits are employed. Remember the secondaries of the transformers should be aligned first and the primaries next unless otherwise indicated by the manufacturer. Proper alignment will be indicated when the response curve is shaped to the form desired and not merely to maximum response.

If the desired response curve cannot be obtained or if the receiver dial does not tune exactly to 1,500 kc at the best response obtained, the receiver tuning dial should be slowly turned, while the necessary oscillator trimmers are retuned for the proper response, until the 1,500-kc point is reached.

After making sure this alignment is correct, set the receiver dial and the frequency dial of the sweep generator to some low frequency, say 600 kc. The sweep-width control of the generator is not usually changed. The next step is to align the oscillator again for the proper response curve, adjusting only the low-frequency trimmers, or padders as they are often called. Next adjust the r-f stages again, in the same order as previously described, for the proper response curve. The low-frequency alignment can be checked for accuracy in the same manner as the high end and readjustments made if necessary. After the low-frequency adjustments are made it may be desired to go back to the high end of the band and check the alignment there again and readjust if necessary.

The foregoing procedure will usually be sufficient for most practical purposes. However, for greater accuracy, when a set is badly aligned to begin with, a point in the middle of the dial may be chosen, and the frequency control of the sweep generator set to this frequency. The oscillator and r-f stages are then checked to see if they are likewise aligned on this middle frequency. If not, readjust the necessary trimmers. After all these steps have been completed, it pays to recheck the low and then the high end again. Middle-frequency alignment points are not usually necessary in visual alignment.

The so-called r-f response of a high-fidelity receiver, obtained by the method indicated, is usually quite different from the over-all i-f response alone, even though the r-f response includes the i-f response. To indicate this the r-f response curve of the receiver from which the over-all i-f curve of Fig. 16-56 was taken is illustrated in Fig. 16-58A. Note that although both curves are triple-peaked, the center peak of the r-f response is higher than its two adjacent peaks whereas, for the i-f response, the opposite is true.

The r-f response curve illustrated in Fig. 16-58A is the correct curve required for the proper alignment of a certain receiver. If this receiver was to have been aligned by the meter method, a maximum meter reading would *normally* indicate proper alignment. However, since the response of this stage has more than one

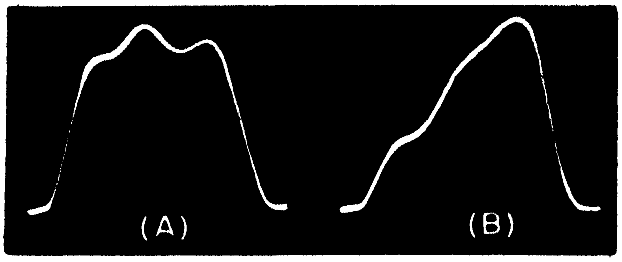


Fig. 16-58.—(A) is the correct r-f response curve of a high-fidelity receiver obtained by the visual method; (B) is an incorrect one, aligned by the meter method and then checked visually.

peak, this maximum output indication may have been found at a wrong alignment point. To prove that this is possible, the same receiver producing the correct response of part (A) of Fig. 16-58 was aligned by the meter method to what was considered correct alignment. Then the over-all r-f response curve of the receiver was checked to see how closely it would resemble that obtained by visual alignment; the resulting pattern appears in Fig. 16-58B. Note the complete absence of symmetry in this curve and how much it differs from the correct response.

VISUAL ALIGNMENT OF F-M RECEIVERS

The greatest difference between a-m and f-m receivers lies in their detector (i.e. demodulation) circuits. Practically all modern a-m receivers employ a diode detector type of circuit whereas in modern f-m receivers, three principal types of detector circuits are found, namely the discriminator detector, the ratio detector, and the locked-in oscillator detector.

In f-m receivers the detector circuit, in conjunction with what is generally referred to as the detector transformer (which is double-tuned), possesses a certain response characteristic by itself. This is a situation different from that in a-m receivers where the detector is used solely for rectification purposes. This f-m detector response should have approximately the same shape no matter what type of detector circuit is used.

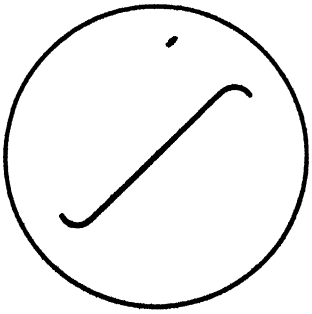


Fig. 16-59.—Ideal response curve of an f-m detector.

An ideal f-m detector response curve is illustrated in Fig. 16-59. Note that it is in the form of an "S". In actual alignment the "S" pattern may appear inverted from that shown but, in whatever position it appears, the detector transformer has to be aligned to produce a response that is equally linear on both sides of the central frequency point of the curve. When this "S" shaped characteristic is to be obtained, the detector is used to actually *demodulate* the input f-m sweep signal.

The different visual methods of r-f, oscillator, and i-f alignment of f-m receivers depends upon the type of detector circuit employed. In the following analyses, three individual f-m receiver alignment procedures will be discussed, one for each different detector circuit.

Alignment of Discriminator Detector

A discriminator detector circuit appears in Fig. 16-60. This circuit is just one of many modifications that may exist. In place of inductance *L* there may be

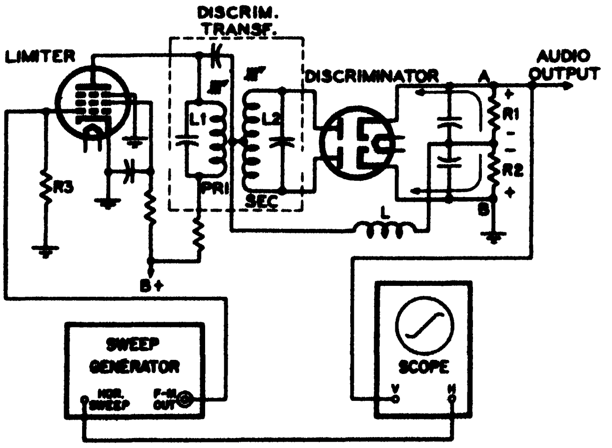


Fig. 16-60.—Circuit arrangement for visually aligning the discriminator detector.

a resistor or no component at all. The output circuit which now consists of two resistors and two filter capacitors may be different from that shown. Whatever modification is used it will have a resistive load in the output circuit similar to resistors *R1* and *R2* in the drawing. It is this resistive load which is important in the alignment process. The circuit illustrated is shown in conjunction with a limiter stage which is supposed to precede this type of detector circuit. When the discriminator transformer is properly tuned and the incoming signal is equal to this resonant frequency, there will be no output across the combination of resistors *R1* and *R2*, that is, between points *A* and *B*, This is so because the currents flowing through the diodes and, hence, through the individual load resistors are equal to each other and opposite in sense, and cause equal but opposite voltage drops across the load resistors. Alignment of the discriminator is based upon this fact. Discriminator alignment means that the transformer is actually being tuned to its proper resonant frequency, which is the same as that of the i.f.'s of the receiver. In the majority of modern receivers this frequency is equal to 10.7 Mc. (Some other uncommon i.f.'s that may be found are 4.5 Mc, 8.3 Mc, and 9.1 Mc.)

In visually aligning this discriminator transformer, the oscilloscope connections are made as indicated in Fig. 16-60. The high side of the vertical input is connected to point *A* and the ground connection made to point *B* or to any similar ground connection of the receiver. The f-m output cable of the sweep generator is connected to the grid circuit of the limiter tube preceding the discriminator as shown in the drawing. The ground side of this cable is connected to any common ground of the receiver. The sine-wave modulating output signal of the sweep generator is fed to the horizontal input of the oscilloscope. The frequency of the sweep generator is set to the i.f. of the receiver. The over-all linear range of the "S" response should be at least 150 kc. The sweep-width control should, therefore, be adjusted to produce a minimum deviation of about 200 kc, to more than cover this minimum required linear range. However, many detector circuits have a linear response greater than 150 kc, so in setting the sweep width, it is suggested that the control be adjusted to the point where at least the upper and lower peaks of the discriminator characteristic are visible. Sweep widths of 300 to 400 kc are commonly used, so that the complete output curve can be visualized.

With these preliminary settings made, some form of "S" curve should appear on the oscilloscope unless the circuit is completely out of alignment. In aligning the discriminator transformer, the primary circuit determines the linearity of the discriminator response. The exact balance or symmetry of the response is determined by the secondary circuit. The primary circuit (*L1* in Fig. 16-60) is adjusted first to obtain the best possible linearity in the "S" curve. Next adjust the secondary of the discriminator transformer (*L2* in Fig. 16-60) to center the i-f point in the middle of the curve. Since the secondary is tuned last there is a possibility that the linearity of the response may be thrown off. If this is noticed in the pattern, readjust the primary of the discriminator transformer for proper linearity.

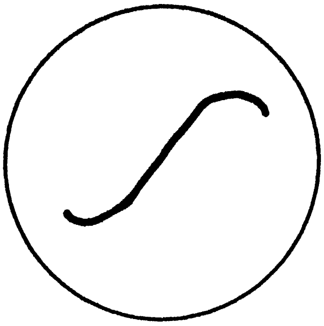


Fig. 16-61.—F-m detector response curve requiring improvement in linearity.

The need for a linearity adjustment is easily deduced from the shape of the curve. For example the hand-drawn curve of Fig. 16-61 illustrates the need for such an adjustment. However, it is quite difficult at times to tell whether or not the secondary adjustment will produce a balance in the system. The need for this balance adjustment can easily be noticed by the use of a marker signal set to the i.f. of the receiver. *When the marker falls in the center of the linear portion of the curve, the secondary is correctly aligned.* Fig. 16-62 illustrates a properly aligned discriminator response with an i-f marker at the center frequency of the curve. The marker generator can also be used to check the linear bandwidth of the response by simply noting at what frequencies the peaks of the curve appear.

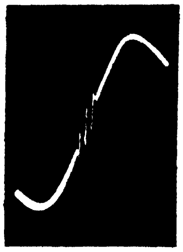


Fig. 16-62.—Properly aligned discriminator response with an i-f marker at its center frequency.

If a wobbulator is used instead of a sine-wave sweep generator a double "S" shaped characteristic similar to that of Fig. 16-63 should appear. For correct alignment the two curves should be mirror images of each other, both should be linear, and the crossover point should be in the center of each curve.

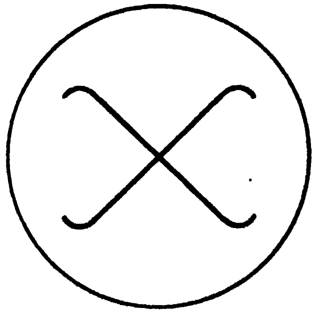


Fig. 16-63.—Properly aligned discriminator response when a wobbulator is used instead of a sweep generator.

Alignment of Limiter and I-F Stages

The alignment of the limiter and i-f stages of the f-m receiver is done in a manner similar to the alignment of a-m receivers. The only difference is that the limiter stage, preceding the detector, acts as a rectifier. Fig. 16-64 illustrates the block diagram needed for visual alignment of the remainder of the receiver. The action of the limiter is such that grid current flows during part of the input signal cycle, and as a result, a current flows through the limiter grid resistor *R3*, putting a potential across the resistor as shown. In other words the grid and cathode of the limiter can be considered as a rectifier with the grid functioning as the plate of the rectifier. The greater the signal input, the greater will be the voltage drop across *R3*, and the greatest signal input to the grid of the limiter will occur when the i-f transformers are correctly aligned.

Based on this limiter action, the oscilloscope is removed from the detector load and placed across the grid resistor of the last limiter stage, if more than one is employed. Some limiter circuits employ two grid resistors, one of them usually much smaller than the other. If this situation exists, it is advisable to place the vertical input terminals across the smaller resistor. In most instances it is advisable to use a large resistor (about 100,000 ohms) in series with the high side of the vertical input of the oscilloscope to prevent the oscilloscope from loading the limiter circuit.

The sweep generator's f-m output cable is placed at the grid input to the i-f amplifier preceding the limiter. The frequency setting remains as before, at the i.f. of the receiver, but the sweep-width control should be increased if necessary to encompass the complete i-f response curve. A sweep width of 600 kc is advisable, or if this is not available from the sweep generator, the greatest possible sweep width should be used. The amplitude control of the sweep signal output should be adjusted to the smallest value which will produce a suitable response curve on the oscilloscope. This preliminary adjustment for a low input signal is primarily intended to prevent saturation of the limiter. If saturation occurs, it will be difficult to tell when the transformer is tuned to resonance.

If two limiter stages are employed (i.e. cascaded limiters), we have to consider the type of coupling between the stages. Slightly different alignment procedures are followed for tuned (transformer) coupling, and for untuned (*R-C* or impedance) coupling. With untuned coupling between the limiter stages, place the oscilloscope in the grid circuit of the first limiter as previously described. The sweep generator's f-m output cable should then be placed at the input to the last i-f stage, as in the case of the single-stage limiter circuit. When the coupling between the two limiter stages is tunable, the oscilloscope is placed in the grid circuit of the second limiter rather than the first. The sweep generator is then placed at the grid input to the first limiter, and the limiter is adjusted as the manufacturer may specify.

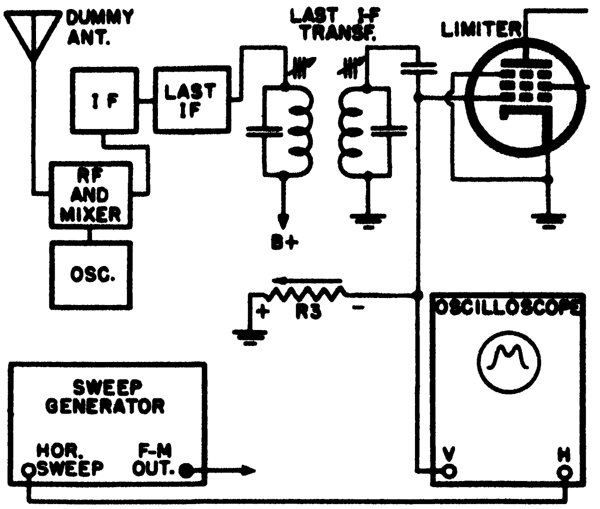


Fig. 16-64.—Block diagram of the circuit needed for visual alignment of the remainder of the f-m receiver.

With these preliminaries completed, the last i-f transformer is aligned first. The sweep signal is inserted at the grid of the last i-f stage. Whatever limiter system is employed, if the coupling of the first transformer to be aligned is below the "critical" point, then only a single-peaked response curve will result. Therefore, the i-f transformer should be adjusted for maximum response at the peak of the curve as seen on the oscilloscope. The peak should normally fall at the center of the curve. The desired pattern is shown in Fig. 16-65A. This is the type of single-peaked pattern usually required by the manufacturers of f-m receivers. If this last transformer stage is overcoupled, the alignment should be made so that the two outside peaks of the curve fall symmetrically on either side of the center frequency dip. The correct pattern is shown in Fig. 16-65B where *a-b* is equal to *b-c*.

The next step is to move only the sweep generator, with the frequency and sweep-width settings intact, to the grid of the preceding i-f tube and align the next i-f transformer starting with the secondary.

When the next to the last i-f transformer has been aligned, the picture on the oscilloscope will be the overall response curve of the last i-f transformer plus the one just aligned. If both i-f transformers are single-peaked, the latter stage should be aligned so that the over-all response is a single-peaked curve similar in shape to that of Fig. 16-65A. Since one curve is effectively added to the other, the amplitude of the sweep signal output should be decreased accordingly. If the stage now just aligned was overcoupled (that is, has a double-peaked response) then no matter how the last i-f transformer is coupled, the correct over-all picture will be a double-peaked response that may or may not be symmetrical about its resonant frequency point. Whatever the case may be, one should know the exact shape of the response to look for. The final curve may contain two peaks that are not too sharply defined, so alignment should be done carefully.

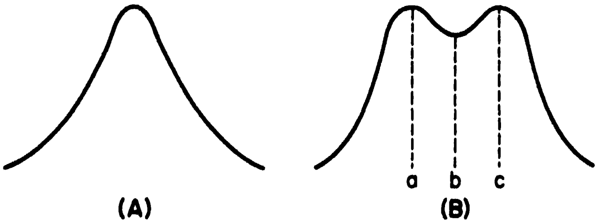


Fig. 16-65.—Desired responses of the last i-f transformer; (A) when this transformer is undercoupled, and (B) when it is overcoupled.

The immediately preceding i-f transformer and the others on back to the first i-f transformer in the set are aligned next, in that order. In each case the f-m output cable of the sweep generator is placed at the grid of the amplifier preceding the i-f transformer to be aligned. In each alignment the picture on the oscilloscope will be a composite response curve of as many i.f.'s as follow the one being aligned. In all cases, whether the i.f.'s are all double-peaked, all single-peaked, or a mixture of both, the picture should resemble as closely as possible a *symmetrical response curve,* unless otherwise noted by the manufacturer's data, either with double peaks that may or may not be too well defined, or just a single-peaked curve. In all these alignments, the output of the signal generator should be as low as possible, so that the limiter is not driven into its saturation or limiting action point.

If individual i-f response curves are desired it is possible to employ a probe detector. In this case the oscilloscope's vertical input is connected to one side of the probe and the other end of the probe connected to the output of the amplifier following the stage being aligned. If frequency checking or bandwidths of the response curves are desired, marker signals can be used in the manner previously outlined.

Oscillator and R-F Alignment in Limiter Discriminator Receivers

To align these stages, the oscilloscope remains across the limiter grid resistor, but the sweep-generator cable is coupled to the antenna terminals of the receiver through the dummy antenna specified by the manufacturer. The sweep-width control of the generator should be advanced if possible to the point where more than the complete response curve is seen. The sweep-generator attachment to the horizontal terminals of the oscilloscope remains as is. Only the frequency setting of the sweep generator is changed. The oscillator and r-f alignment procedure as discussed under a-m visual alignment is now applied here. The primary difference is that the high and low frequencies of the f-m band used for this alignment process should be approximately 105 Mc and 90 Mc respectively. The adjustments of the r-f and oscillator sections should be for maximum and symmetrical response curves unless otherwise noted by previous i-f alignment curves or by manufacturer's requirements.

Ratio Detector and I-F Alignment

The second important type of detector circuit employed in f-m receivers is the ratio detector. When this detector is used, it may or may not be preceded by a limiter stage. In fact in many cases it is difficult to tell whether or not the stage preceding the ratio detector is a limiter or just another i-f amplifier. Actually, the only major difference is that grid current always flows in the limiter stage. The original ratio-detector design was intended for use without a limiter stage. Whatever the case may be, once again we say that the manufacturer's visual-alignment instructions should be followed, if any are given.

The ratio-detector circuit is similar to the discriminator detector; a duo-diode tube is generally used but, in the ratio-detector circuit, the diodes are wired in series aiding instead of series opposing, as in the discriminator circuit. There are many modifications of the ratio-detector circuit; the one shown in Fig. 16-66 is one of the original types. As in the discriminator, circuit inductance *L* may not exist at all or a resistor may be in its place, and the output load circuit may appear different from that shown. Whatever circuit is found in the f-m receiver to be aligned, the arrangement of Fig. 16-66, in conjunction with the discussion to follow, will allow for the alignment of all types.

The load resistance *R1* may be a single resistor as indicated in Fig. 16-66 or it may consist of two equal

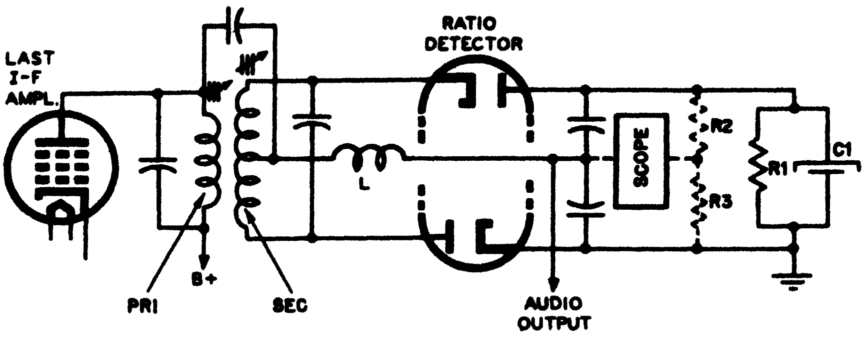


Fig. 16-66.—Schematic of a ratio detector circuit.

resistors as shown by the dashed resistors *R2* and *R3* in the circuit. If only one resistor *R1* exists, then two resistors of about 100,000 ohms each with a tolerance of ±5% can be soldered across *R1* to simulate the circuit including *R2* and *R3*. In the alignment of this stage the vertical terminals of the oscilloscope are placed as indicated in Fig. 16-66 or between the audio output point and ground. The modulating sine-wave output from the sweep generator is fed to the horizontal terminals of the oscilloscope. The f-m output cable of the sweep generator is connected to the grid of the stage preceding the ratio detector. The frequency of the sweep generator is set to the i.f. of the receiver and the sweep-width control is properly adjusted, 300 to 400 kc being an average width to use. Several variations in the visual alignment of the ratio detector and i-f stages are specified by different manufacturers. Three different methods will now be described.

In one method the tuning component in the secondary of the ratio-detector transformer is turned to its minimum. The primary of the transformer is then aligned for a single-peaked i-f curve of maximum response similar to that shown in Fig. 16-65A. The sweep generator is then moved back a stage and the next to the last i-f transformer is aligned for the proper pattern. The procedure from this point onward exactly follows that of the i-f alignment of the discriminator detector type of f-m receiver, including the remarks about double-peaked i.f.'s. *After* all the i.f.'s are aligned, the discriminator secondary is adjusted to give the proper S-shaped curve. The f-m output cable of the sweep generator for this last adjustment can be placed either at the grid of the last i-f tube or left at the r-f grid of the converter tube. Whichever is used, remember to control the amplitude of the f-m output signal from the sweep generator.

A second method involves aligning the complete ratio-detector transformer first. The oscilloscope is placed in either of the positions mentioned in the previous method and the secondary of the ratio-detector transformer is completely detuned. The f-m output cable of the sweep generator is placed at the grid of the last i-f amplifier with the same settings as in all the detector visual alignments thus far described. The primary of the detector transformer is then tuned for the proper single-peaked curve similar to that of Fig. 16-65A. Then the secondary of the detector transformer is aligned to give a symmetrical S-shaped curve. The ratio detector is now completely aligned. A marker generator can be used to check the midfrequency of the detector response. The alignment of the i.f.'s is based, in this case, not on the picture of the i-f response curves, but rather on the detector output. The signal generator is gradually moved back as in the other i-f alignments and the proper i.f.'s adjusted. The adjustment in each case is for a continued symmetrical and linear S-shaped response curve of maximum amplitude.

A third method involves opening up the hot lead of electrolytic capacitor *C1* in Fig. 16-66, so that the circuit functions as an a-m demodulator. The vertical terminals of the oscilloscope are then placed across load resistor *R1*. The sweep generator is set at the same frequency and sweep width, and is moved one stage backward each time as the alignment progresses. Some manufacturers specify removal of the first audio tube. Alignment starts with the complete detector transformer and continues backward to the first i-f transformer. If the i.f.'s and detector transformer are all single-peaked, the alignment should follow the previous i-f alignment where the i-f transformers are adjusted for a symmetrical single-peaked resonance curve. Care should be taken in tuning overcoupled i.f.'s for the double-peaked curve; in some instances, it may be advisable to detune the secondary of overcoupled i.f.'s, tune the primary for a single-peaked curve, and then tune the secondary for the double-peaked curve. Remember that when aligning some of the first and perhaps second i-f stages the response curve, which may include overcoupled i.f.'s, may not contain any evidence of double peaks due to the effective combination of all the i-f resonance curves. The over-all i-f curve may contain only a single peak, but it will be a broad response curve.

After this alignment is completed leave the sweep generator as is and capacitor *C1* of Fig. 16-66 is reconnected to its former position. The vertical terminals of the oscilloscope are then placed as indicated in the drawing of Fig. 16-66 or across the point of audio output to ground. With these connections made, the oscilloscope should show a symmetrical and linear S-shaped curve, because the detector tube is once again functioning as an f-m demodulator.

Oscillator and R-F Alignment of Ratio-Detector Receivers

Visually aligning the oscillator and r-f stages is very simple. The adjustments and preliminary settings at the front end of the set and the sweep generator remain essentially the same as for the discriminator type of f.m. receiver, but the positioning of the oscilloscope may vary. For example, for the case where capacitor *C1* is opened, the oscillator and r-f stages can be aligned for maximum and symmetrical i-f patterns on the oscilloscope with the oscilloscope placed across *R1* in Fig. 16-66. Then the detector circuit could be checked for the proper "S" characteristic with the capacitor in place. Another method would be to keep the oscilloscope in the position as shown in Fig. 16-66 or across the audio output and adjust the oscillator and r-f circuits while the secondary of the ratio-detector transformer is detuned. The adjustments are for the proper i-f response curves. Afterward the detector can be aligned. In this method the detector transformer is actually aligned last in the complete receiver system.

Still another method would be to connect the oscilloscope in either manner mentioned in the latter method and adjust the oscillator and r-f stages after all the other alignments have been made. The adjustment would be for a continued maximum and symmetrical S-shaped response curve from the detector.

If the tube preceding the ratio detector is definitely a limiter, (i.e. grid current flows and a negative bias is developed across the grid resistor) then the alignment of the receiver could be accomplished in another manner. The ratio-detector circuit should be completely aligned *first* by any method just outlined. After this the oscilloscope is placed in the limiter grid circuit and the remainder of the receiver's alignment carried out in the same manner as receivers employing discriminator detectors.

Visual Alignment of the Locked-In Oscillator Detector Receiver

The third principal type of f-m detector, namely the locked-in oscillator detector, as far as is known, is used by only one manufacturer. Consequently this detector is not found to the same extent as the other f-m detectors but it is, nevertheless, important enough to be included. This circuit employs a special pentagrid tube which functions as the locked-in oscillator and detector. A schematic of this f-m detector circuit as found in a typical receiver appears in Fig. 16-67.

The network consists of three different tuned circuits, *L1-C1, L2-C2,* and *L3-C3*. Each one of these tuned circuits is resonant to the same frequency, namely the i.f. of the receiver. The input transformer (that containing *L1-C1*) is considered as the last i-f transformer of the receiver. The *L2-C2* tuned system in conjunction with the rest of the circuit functions as a Colpitts oscillator. The *L3-C3* tuned arrangement is known as a quadrature circuit. The remainder of the i-f and r-f sections of such f-m receivers are similar to the other types of f-m receivers.

The visual alignment discussed here is carried out on an actual receiver — Philco model 48-482. The i.f.'s are aligned first for the correct response curve and the detector afterward. The visual-alignment procedure is as follows. The high side of the vertical input of the oscilloscope is connected to the junction of *R1* and *C3* as shown in Fig. 16-67. The ground lead of the oscilloscope is connected to the chassis of the receiver. It is across resistor *R1* that the audio-frequency output appears. For the i-f alignment the oscillator grid, the first grid of the detector tube, is grounded so that the oscillator section of the tube is made inoperative. When this is done the tube will no longer be effective as an f-m detector but will function as an a-m grid-leak detector.

The frequency control of the sweep generator is set to the i.f. of the receiver, which in this case is 9.1 Mc. The f-m output cable is connected to the grid of the last i-f amplifier and the sweep-width control is set to a large deviation (at least 400 kc). The oscilloscope remains where it is for the complete receiver alignment. The primary of the last i-f transformer is aligned for a symmetrical single-peaked maximum response on the oscilloscope screen. The rest of the i.f.'s are similarly aligned with the output cable of the sweep generator being moved back each time and placed at the grid of the i-f amplifier preceding the i-f transformer to be aligned.

After the i.f.'s are aligned, the ground is removed from the oscillator grid. A short is then placed across

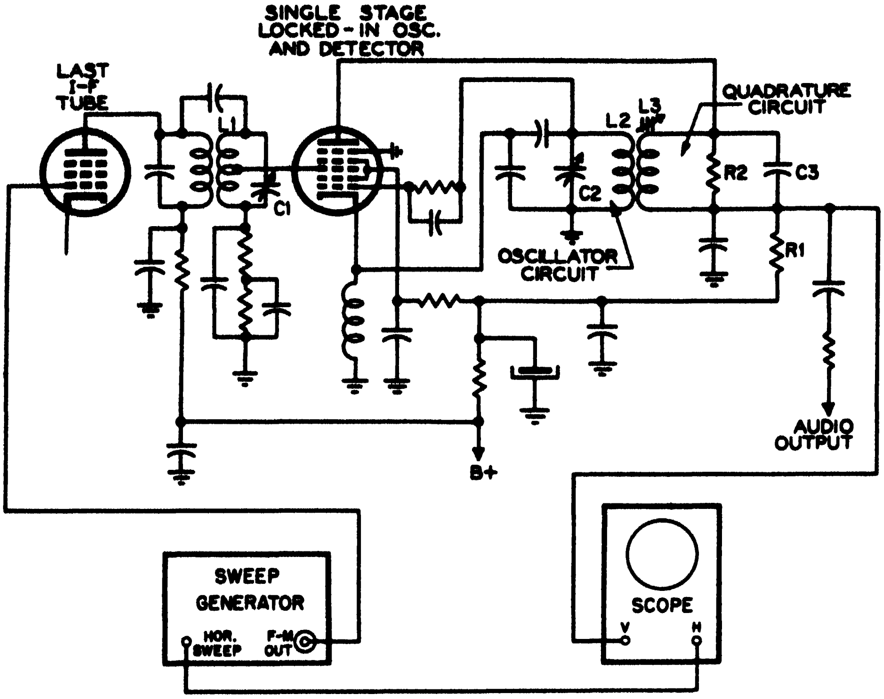


Fig. 16-67.—Schematic of a locked-in oscillator detector, including oscilloscope and sweep generator setup for visual alignment.

the quadrature network (across either component *L3*, *R2*, or *C3*, whichever is easier to get at), and the f-m output cable of the sweep generator remains connected as in the very last step. The frequency of the sweep generator is set at 9.1 Mc, ***unmodulated***, or if this is not possible, a 9.1-Mc unmodulated signal should be used from a marker generator. The oscillator trimmer, *C2* in Fig. 16-67 is then adjusted for a zero beat, as indicated by a minimum signal on the oscilloscope and in the loudspeaker of the receiver.

After the oscillator alignment, the short across the quadrature circuit is removed for alignment of this section, but the sweep generator remains where it is. The generator is frequency-modulated at a center frequency of 9.1 Mc with a minimum sweep width of 200 kc. The quadrature coil *L3* (see Fig. 16-67) is then adjusted for linear detector response on the oscilloscope screen as indicated in Fig. 16-68B. This linear response was taken when the sweep width was exactly 200 kc. From this curve we see that the linear part covers more than the minimum required bandwidth of 150 kc. A 9.1-Mc marker can be used to determine whether or not the response at midpoint of the linear curve is at the i.f. A typical response curve when the quadrature circuit is misaligned is illustrated in Fig. 16-68A when the sweep width is set at 200 kc. The misaligned response curve shown at Fig. 16-68C is taken at the detector output when the quadrature network and also the secondary of the last i-f transformer are both misaligned. This latter misalignment is also under the condition of a 200-kc sweep width.

To align the high-frequency oscillator and r-f section visually, the f-m output cable of the sweep generator is connected to the antenna input terminals or to the dummy antenna recommended by the manufacturer

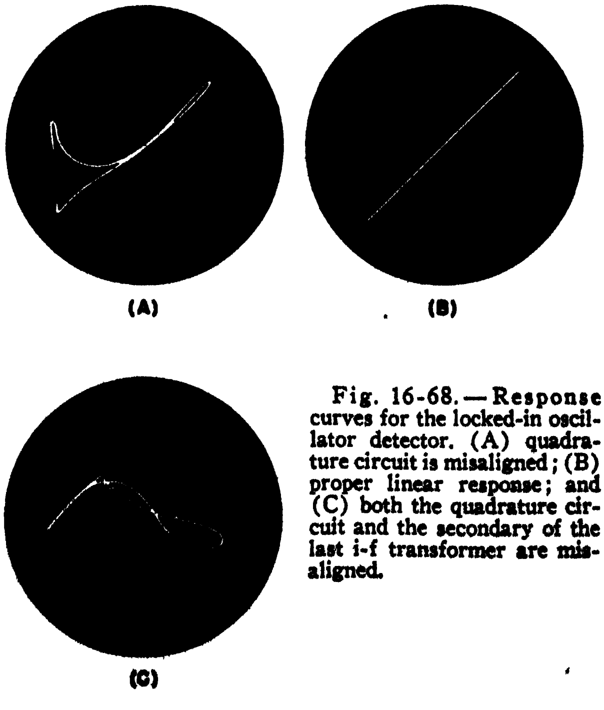

(A) (B) (C)

Fig. 16-68.—Response curves for the locked-in oscillator detector. (A) quadrature circuit is misaligned; (B) proper linear response; and (C) both the quadrature circuit and the secondary of the last i-f transformer are misaligned.

and the oscilloscope remains where it is. The frequency control of the sweep generator and receiver dial are first both set at the same frequency on the high end of the band for the capacitor trimmer adjustments, and then at the low end for the coil adjustments. Use high and low frequencies of 105 Mc and 90 Mc respectively if none are recommended. The sweep-width control of the sweep generator is set wide enough to give a good picture of the over-all band-pass characteristic. The oscillator and r-f stages can then be aligned in proper order for maximum linear detector response on the oscilloscope. Marker signals can be used during this alignment procedure to determine the midfrequency of the response.

An alternate method of aligning the r-f and oscillator sections would be for a maximum and symmetrical response of the i-f stages. To do this the grid of the oscillator section of the detector tube has to be grounded. With this method it is possible that the oscillator and r-f be aligned immediately after the i-f alignment, and the quadrature circuit of the f-m detector aligned last.

VISUAL ALIGNMENT OF TELEVISION RECEIVERS

We now come to the type of receiver where visual alignment is most important. In alignment of a-m and f-m receivers it is possible to use a complete meter method (although the visual method is far more accurate) but for television receivers this is not so. Visual alignment of the r-f and video i-f sections of these receivers is a necessity because of the specially shaped curves required. Since visual alignment of part of this receiver is required, it is a simple matter to visually align the complete receiver including the sound section. It is important to remember that the marker and sweep generators used for alignment of television receivers must have a frequency range that includes the r.f.'s and i.f.'s of the receiver.

There are five main sections of a television receiver that have to be considered in the process of alignment; these are the sound section, the trap circuits (if any are employed), the video i-f section, the oscillator circuits, and the r-f stages. The sound section of television receivers is very much the same as the circuits of f-m receivers. Since this is so and since most manufacturers suggest aligning the sound section first, we will start our discussion with this part of the television receiver.

The Sound Section

Sound i-f amplifiers, limiters (if any are used), f-m detector, and the audio system represent the order of the stages in the sound section of television receivers. This is exactly the same as in f-m receivers. The number of sound i-f amplifiers used in television receivers may vary according to the make of the set and whether or not the intercarrier sound system is used in the receiver. As far as is known only discriminator detectors or ratio detectors are used as f-m demodulators in the sound section of television receivers.

The chief difference between standard f-m broadcasting and the f-m sound part of television broadcasting is that, in the former, ±75 kc deviation of the transmitted f-m signal is considered 100 per cent modulation and, in the latter, it is only ±25 kc deviation. This means that the bandwidth of the tuned circuits in the sound section of television receivers need not necessarily be as broad as those in f-m receivers. However, in order to prevent oscillator drift from causing the f-m i-f signal of television receivers to fall outside the response of the sound i-f, the tuned circuits are normally made much broader than necessary. In some cases they may be broader than those of f-m receivers to compensate for this frequency drift. The sweep-width control of the sweep generator should be adjusted to produce somewhat more than the complete response curve.

The placement of the oscilloscope and its connections to the f-m detector circuit and sweep generator for visual alignment of the sound section of television receivers is exactly the same as for f-m receiver alignment. The complete visual-alignment procedure of this sound system follows any of those described for the i-f and detector sections of f-m receivers. It is not necessary to repeat this alignment here. Whatever type of f-m detector is used in the sound system, refer to the proper section under f-m receiver alignment and follow the methods outlined there.

The important thing to remember about alignment of this section is the intermediate frequency employed. If the television receiver is of the intercarrier type, the final sound i-f signal of any receiver will always be 4.5 Mc. This 4.5 Mc sound i.f. is the result of mixing the video and sound i-f signals from the front-end converter in the video detector. One of the signals that results from this mixing process is a 4.5-Mc signal, frequency modulated with the sound intelligence. This 4.5-Mc signal is fed to the grid of the first sound i-f amplifier through some tuned circuit arrangement which is resonant to 4.5 Mc. This arrangement may be a single series- or parallel-tuned circuit or a double parallel-tuned one, and may appear at the output of the video detector or at one of the video amplifiers or at the input to the first sound i-f amplifier itself. Whatever type of tuned circuit is used, it is generally referred to as the 4.5-Mc *sound take-off* circuit, transformer, or trap, or the *intercarrier take-off trap.*

In receivers other than the intercarrier type, which receivers we will refer to as the conventional type, the resonant frequency of the sound i-f transformers is the same as the sound i-f output from the front-end converter tube. The exact frequency of this sound i-f signal is not always the same as it is for intercarrier receivers but differs for different model receivers. The i.f.'s in question may have a value anywhere from 20 to 40 Mc. The exact i.f. used is usually mentioned by the manufacturer of the receiver. The tuned circuit preceding the first sound i-f amplifier in such receivers may appear in the output of the front-end converter tube or in the output of the first or second video i-f amplifiers. Whatever type of receiver is being aligned the frequency control of the sweep generator should be set to the correct sound i-f of the receiver.

Alignment of First Sound I-F Tuned Circuit

Visual alignment of the sound section of either type of receiver is, as mentioned previously, carried out in the same manner as for f-m receivers. The only difference that should be noted in this section is the placement of the f-m output cable of the sweep generator for alignment of the first i-f tuned circuit or transformer.

For intercarrier type of receivers where the 4.5 Mc sound take-off circuit appears after a stage or two of video amplification, the 4.5-Mc f-m signal output from the sweep generator should be fed to the grid circuit of the video amplifier that immediately *precedes* the sound take-off circuit. If the sound take-off circuit appears immediately after the video detector, the output cable of the sweep generator is usually placed at the electrode of the detector tube to which the take-off circuit is effectively connected, preferably through a 0.001 μf capacitor. In placing the sweep generator for alignment of this stage, it is important that the generator itself not load down the take-off circuit to the point where it will cause erroneous alignment.

Wherever the 4.5-Mc take-off circuit is employed, it should be aligned as just another i-f stage. The oscilloscope remains where it is after the regular i-f alignment and the 4.5-Mc take-off circuit is aligned for the proper response curve using the same sweep width as for the regular i-f alignment. Remember that the resulting curve on the oscilloscope is the over-all response of all the i-f stages as well as the take-off circuit. Some manufacturers suggest placing the sweep generator before the take-off circuit for alignment of the complete f-m sound section. This method is all right if the sound section is not badly out of alignment.

Visual alignment of the tuned circuit or transformer preceding the first sound i-f amplifier in conventional type television receivers is carried out in a manner similar to that just described. The f-m output cable of the sweep generator is placed at the grid input of the stage preceding the first sound i-f tuned circuit, whether this stage be the front-end converter tube or the first or second video i-f amplifier. Wherever this first tuned i-f circuit appears, the same frequency and sweep-width control adjustments of the sweep generator that were made for alignment of the other i-f stages are also used for alignment of this circuit.

I-F Trap Circuits

Receivers that do not employ the intercarrier system contain a number of traps within the *video* section of the receiver to suppress certain frequencies that would normally cause interference of the reproduced picture. The traps employed are either series- or parallel-tuned circuits that are sharply resonant. The trap either absorbs energy or causes a loss in gain at its resonant frequency.

These traps, although they help shape the over-all video i-f response curve, are essentially inserted to prevent interference from undesired sound and video i-f signals. The sound i-f signal of the receiver is one type of interfering signal. This i-f signal is referred to as the *accompanying sound i.f.* because it is a signal within the channel to which the receiver is tuned. It is also possible to have interference of sound and video signals from adjacent channels which may force their way into the front end of the receiver. If this happens, undesired sound and video i-f signals will appear at the output of the converter tube. These i-f signals, which are commonly referred to as *adjacent* sound and video signals, can easily force their way through the video i-f circuits because they fall on the skirts of the response curve. The adjacent video i-f signal is 6 Mc away from the desired video i-f signal and the adjacent sound i-f signal is 6 Mc away from the accompanying sound i-f signal or 1.5 Mc higher than the desired video i-f signal. The traps are tuned to the frequencies of these interfering i-f signals.

Not all television receivers employ traps for each one of the possible interfering i-f signals; some receivers use more than one trap tuned to the same frequency. In order to present a fair idea of where these undesired frequencies lie on the response curve of the video i-f circuits and to show how they help shape the

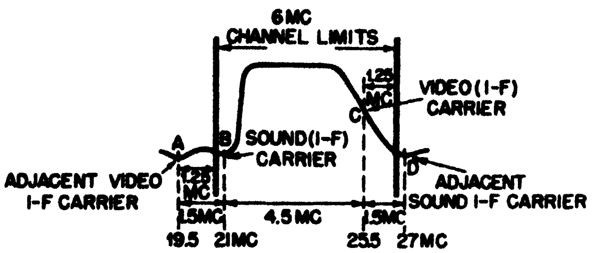


Fig. 16-69.—Ideal over-all video i-f response curve of a television receiver.

curve, we have included an ideal over-all video i-f response curve in Fig. 16-69. In actual practice the shape of the over-all video i-f response curve will vary from that shown according to the type of coupling between the video i-f stages and the number of different traps employed.

The frequencies indicated on the drawing are from a receiver that has a desired video i-f carrier of 25.5 Mc, point *C*. The sound i-f carrier, point *B*, being 4.5 Mc less, has a frequency of 21 Mc. The adjacent video and sound i-f carriers, points *A* and *D* respectively, therefore, have frequencies of 19.5 Mc and 27 Mc, respectively. The video i-f interfering signal comes from the upper adjacent channel and the adjacent sound i-f interfering signal comes from the lower adjacent channel. This is all predicated upon the assumption that the desired channel and either adjacent channel do not have any frequency separation between them.

There are some places where such traps are commonly found. In many cases the traps are located in the plate circuits of the different video i-f amplifiers. At other times they will be found somewhere in the coupling between the video i-f amplifiers. In still other receivers, these traps are found in the cathode circuits of the video i-f amplifiers. Wherever they are found, their basic action is still the same as previously described.

Many of the conventional type receivers employ a 4.5-Mc trap somewhere between the output of the video detector and the grid of the picture tube. This trap is inserted to prevent any possible 4.5-Mc beat-note signal that may appear in the output of the detector from also appearing at the input to the picture tube. This trap is also sharply tuned and has to be aligned to its proper resonant frequency, as well as the other traps.

I-F Trap Alignment

Many television receiver manufacturers suggest aligning these traps with a vacuum-tube voltmeter instead of an oscilloscope. The use of such a voltmeter does not make the process any faster than when an oscilloscope is used. In fact an oscilloscope would be preferable because it has to be used for alignment of the video i-f response anyway. In most cases the oscilloscope or vacuum-tube voltmeter is placed at the same point no matter which is used. This point is usually across the load of the video detector or of the agc rectifier of the receiver or at some point after either of these. Whatever kind of indicating device is used, in practically all cases an *a-m signal* (usually audio-modulated) is employed for alignment of these traps. The frequency of the a-m signal is set to the resonant frequency of the trap to be aligned. In most cases the trap is tuned to produce a minimum indication on the meter or oscilloscope.

The reason the indicating device is placed after a diode detector of the system is to have this diode demodulate the a-m signal. The resulting output signal, which is usually audio, will appear on the oscilloscope as a sine wave, and will only cause a deflection on the vacuum-tube voltmeter. Whatever device is used as the source of a-m signal, it should be accurately calibrated because the traps are sharply tuned. If the marker generator has provision for a-m, then it should be used. Some manufacturers suggest lining up the traps during the alignment of the video i-f circuit while others recommend their alignment as a separate procedure. In this section, trap alignment is being considered as an individual part of the alignment procedure of the television receiver.

The input impedance of the oscilloscope is such that when it is placed across the video detector or agc rectifier load, it may cause appreciable change in the effective load on the tube. If this happens it is suggested that a resistor of about 100,000 ohms be inserted between the high side of the vertical input to the oscilloscope and the load of the tube in question.

There are certain prealignment adjustments recommended before alignment of the traps and video i-f circuits. These adjustments are definitely not the same for all types of television receivers. They vary with the type of indicating device to be used, where it is to be placed, the type of circuits employed, and the general design of the receiver. A representative set of prealignment adjustments are as follows. The high-frequency oscillator and r-f sections are disabled by either removing the front end tubes or by tuning to an unused channel. The contrast control is usually placed in the center of its range, or in a total clockwise position, and remains there. A fixed bias is applied to the tubes that normally have agc bias. This bias is necessary to simulate an operating agc voltage. This is also necessary for video i-f alignment as will soon be seen. There are various ways of applying this bias voltage and they depend upon the circuit design of the television receiver. It is beyond the scope of this section to list all the different methods of bias hookup. The exact bias necessary and the manner of applying it are normally given in the manufacturer's service notes that accompany the television receiver.

Once all these prealignment adjustments are made, it is a simple matter to align the traps. The a-m input signal lead can be connected in either of two ways. One method is to attach it to the grid of the tube preceding the first trap and to keep it there for all of the trap adjustments. The other method, and the one usually recommended if there is enough signal amplification, is to place the a-m input signal only at the grid of the amplifier preceding the trap or traps to be aligned.

Whichever method is used, the input signal (amplitude-modulated at about 30 per cent) should be set to the *required* resonant frequency of the trap to be aligned. The trap is then adjusted to produce, on the oscilloscope, a sine wave of *minimum* amplitude. A minimum signal is required because, when the trap is properly tuned, it will suppress the incoming signal to the best of its ability and hence only the weakest signal passes through. Every time a different type of trap is aligned, the generator should be set to that trap's proper frequency and the trap aligned for a similar minimum signal output. *In the alignment of i-f traps the linear sweep of the oscilloscope is used as the horizontal-deflection voltage.*

4.5-Mc Trap Adjustment in Conventional Receivers

We mentioned previously that in many conventional television receivers a 4.5-Mc trap or two may be employed somewhere between the output of the video detector and input to the picture tube. In most cases a single trap is used and it usually appears between the first and second video amplifiers or between the last video amplifier and the input to the picture tube.

Since the trap should be resonant to 4.5 Mc, a signal of this frequency is needed to align it. This a-m signal, after passing through the trap, must be detected for most service oscilloscopes to respond to it. Since the trap appears after the video detector of the receiver, a *probe detector* must be used somewhere after the trap. In many cases the probe detector is inserted between the oscilloscope and the terminal leading to the input of the picture tube. If the trap appears between two stages of video amplification, it may be possible to connect the probe detector to the output circuit of the tube following the trap.

The output cable of the a-m generator should be connected to the grid of the first video amplifier if the trap appears after this tube, which is usually the case. If the trap appears between the video detector and first video amplifier then the a-m signal should be connected somewhere before the trap and the output of the video detector. Care should be taken that the generator does not cause any undesired loading effects on the trap because this may cause erroneous alignment.

After the proper oscilloscope, probe detector, and generator connections have been made, the trap is then aligned to produce a minimum sine-wave signal on the oscilloscope. This is similar to the alignment of the i-f traps. Do not increase the output of the generator to the point where it may cause overloading of the video amplifier in question. Also since we are tuning for a minimum signal on the oscilloscope the output of the generator should not be kept too low because the oscilloscope may not register any pattern when the trap is tuned through its resonant frequency.

Video I-F Stages

We now come to the most important part of the television receiver that requires alignment, namely, the video i-f system. The number of video i-f amplifiers and hence the number of tuned coupling circuits is not the same for each television receiver. The type of coupling also varies with the different models. Single-peaked, stagger-tuned, and over-coupled i-f circuits are the types most commonly employed in television receivers. These circuits can be either single or double tuned. In many stagger-tuned circuits each of the interstage coupling circuits is tuned to a different frequency. In most overcoupled circuits, the individual stages are generally all tuned to the same frequency. However, the coupling between the primary and secondary of each interstage i-f transformer is great enough to produce the desired bandwidth.

Visual alignment of these stages can be performed by using the video detector of the receiver, to which the oscilloscope is connected, or a probe detector for alignment of each stage individually. The sweep generator employed should be capable of a 10-Mc sweep width and also a frequency range to cover all the i.f.'s of the receiver being aligned. An accurately calibrated marker generator is also required, capable of covering the frequencies of the complete i-f circuit being aligned. Whether a probe detector is used or not, in the final analysis, the over-all video i-f response, which includes the response of all the i-f tuned circuits, must be checked on the oscilloscope. For this check, the marker

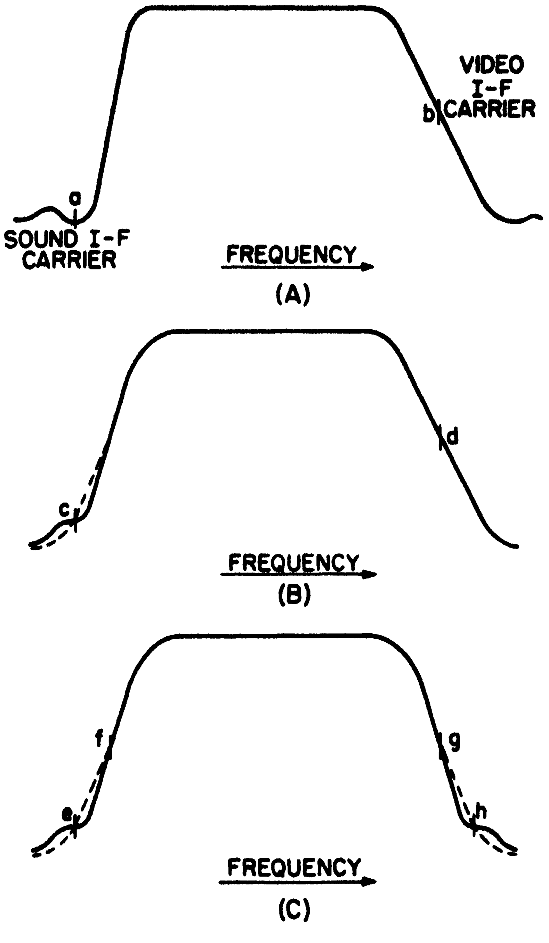


Fig. 16-70.—Ideal over-all video i-f response curve: (A) for the conventional type of receiver; (B) for the intercarrier type of receiver; and (C) for the intercarrier type which tunes above and below the incoming signal.

generator is definitely needed. It is the final shape of this over-all video i-f response which determines whether or not the alignment is correct.

The ideal shape of this over-all video i-f response for the conventional receiver has already been indicated in Fig. 16-69 and it is repeated in part (A) of Fig. 16-70 with only the notations that are of interest to us at the moment. The right slope must be fairly broad compared to the left slope which should be fairly steep. The video i-f carrier should fall approximately in the center of the right slope as indicated by point *b* in the drawing. In other words the response at point *b* should be about 50 per cent of the average maximum response of the curve. Some manufacturers specify that if the video i-f carrier falls along the right slope anywhere between 40 and 60 per cent of maximum response, receiver performance will be acceptable. Point *a* on the left side of the curve of Fig. 16-70A represents the carrier of the accompanying sound i.f. and there should be negligible response at this frequency.

Part (B) of Fig. 16-70 represents an ideal over-all video i-f curve of the intercarrier type of television receiver. The right side of this curve should be the same as that of the conventional type of receiver. Point *d* represents the video i-f carrier and it occurs at a response point of about 50 per cent of maximum. In intercarrier receivers, a certain amount of sound i-f signal must come through. The step on the left side of the response *ideally* indicates the shape desired, where point *c* represents the sound i-f carrier. However, the majority of television receivers simply have the left side of the response shaped as indicated by the dashed line in Fig. 16-70B. Although the former shape is exactly what is desired, the other shape, which also allows a certain amount of sound i-f signal through, is used because it is simpler to obtain in practice.

In intercarrier television receivers where the high-frequency local oscillator is tuned above and below the incoming television signal (for the low and high channels respectively) a different type of over-all video i-f response is required. The ideal curve is shown in Fig. 16-70C by the solid line. Points *e* and *g* indicate the sound and video i-f carriers respectively when the oscillator is tuned above the incoming television signal and points *f* and *h* are the respective video and sound i-f carriers when the oscillator is tuned below the incoming television signal. Video carrier points *f* and *g* should be at approximately 50 per cent of the maximum response. The sound carriers *e* and *h* should fall approximately in the center of the steps on the sides of the drawing. The dashed lines on both sides of the drawing of Fig. 16-70C indicate the shape generally used in practice because of the difficulty in obtaining the steps on both sides of the curve.

In intercarrier receivers the exact response to the sound i-f carrier [see parts (B) and (C) of Fig. 16-70] will be found to vary whether or not a *step* formation is used. There will be found cases where the response suggested is almost 20 per cent of maximum and at other times as little as 5 per cent. As we have always mentioned, the alignment procedure recommended by the manufacturer should always be followed. However, it is important to point out in this section that too high a sound i-f response in intercarrier receivers, especially those that do not shape the response somewhat in step fashion, usually causes a distorted sound output. In most cases this sound i-f response should not be more than about 10 per cent of maximum.

Of course, if an intercarrier receiver is specially designed to produce a certain shape response curve, very little can be done to change the response of the left side, unless perhaps special traps are already employed in the receiver. In the following analysis of the suggested methods of video i-f alignment, the shape of the over-all video i-f response will be discussed last.

Alignment of Video I-F Stages

The exact methods of aligning the video i-f stages vary quite a good deal for different receiver models even though, in the end, they all require over-all video i-f responses that are somewhat similar to each other. Before the alignment procedure is to be started, the required prealignment adjustments should be made. One very important adjustment, as mentioned previously, is the use of a fixed bias on the tubes that normally have agc voltage on them. It is usually suggested that the bias be the same as the value of agc voltage on the tubes when the receiver is in normal operation. The exact bias values and the method of obtaining them usually can be found in the manufacturer's service notes. In some cases the necessary bias is obtained from within the receiver itself and in other cases a bias battery is used and agc rectification is made inactive by removal of the agc rectifier or agc amplifier tube, if one is employed. The high-frequency oscillator and front-end tuned circuits are effectively made inoperative by tuning the channel switch to a "dead spot" on the dial (such as between two channels, between channels 12 and 13 if possible, or to an unoccupied channel) or by removal of the oscillator tube.

Since the required positioning of the contrast control will vary for different receiver models, it is difficult to make one comment about a prealignment adjustment of this control that will hold true for all of them. If the control regulates the bias on the video i-f stages, then it should remain in the extreme counterclockwise position or at a point where it will not cause overloading of any of the tubes. On the other hand, if the control appears somewhere after video detection, its regulation may not be as rigorously determined as in the other case. If no regulation of this control is suggested by the manufacturer, it should be set at its midposition.

The alignment proper of the video i-f stages usually starts with the last i-f stage and works backward. There are some manufacturers who suggest aligning the first i-f transformer first (that at the output of the front-end converter tube) because of special circuit conditions. Our analysis will be based on the former method because it is the one most often used.

As mentioned in the alignment of the trap circuits, the video-detector circuit is, in practically all cases, a single diode detector and hence the vertical input of the oscilloscope is effectively placed across the diode load. If the video i-f signal is fed to the plate of the detector, the load will be in the cathode circuit and vice versa. In the majority of cases the input video i-f signal is fed to the cathode of the video detector and the load appears in the plate circuit. Care should be taken lest the oscilloscope change the detector load by any appreciable amount. If this appears to be happening, insert a 100,000-ohm resistor in series with the high side of the vertical input of the oscilloscope. In Fig. 16-71

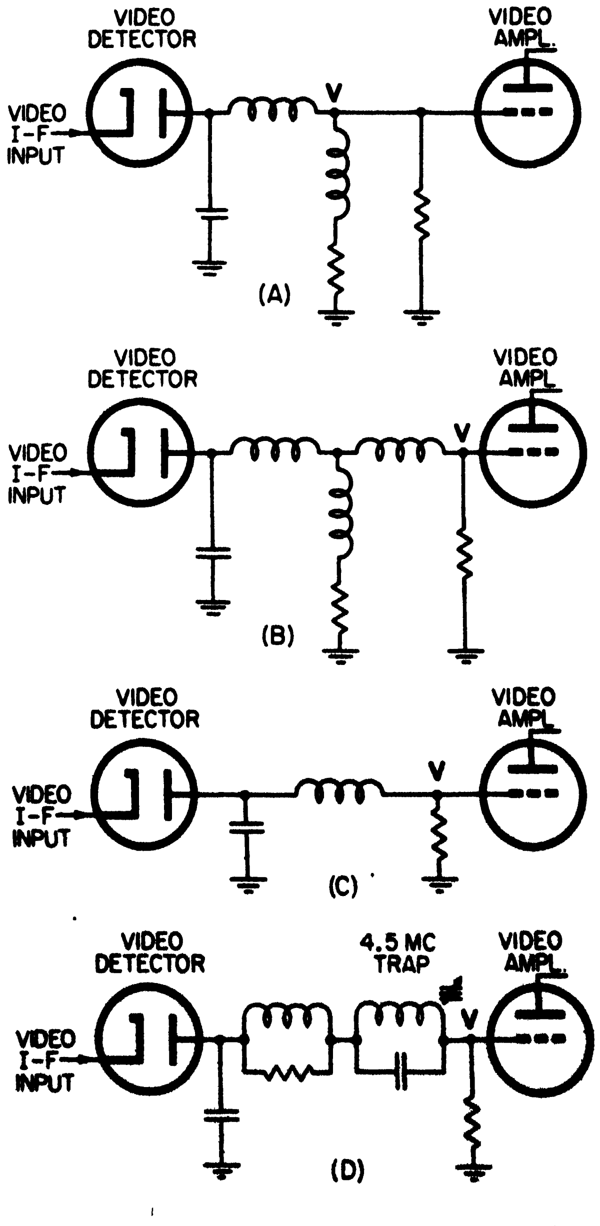
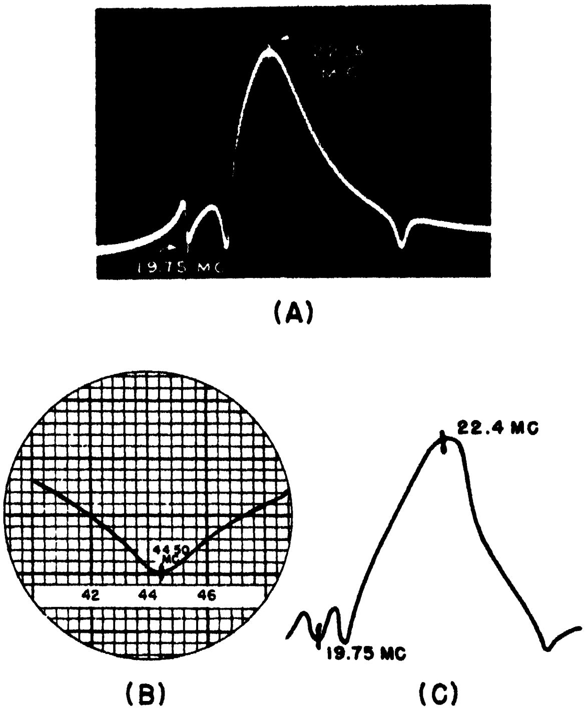


Fig. 16-71.—Four typical video detector circuits.

appear four typical video-detector circuits. In each circuit the video i-f signal is shown fed to the cathode of the tube. The detector load appears in the plate circuit of each tube. Each loading network is different from the others. The coils in each plate circuit function as high-frequency compensators (see Chapter 10). The tunable *L-C* circuit of part (D) represents a typical 4.5-Mc trap as used in conventional television receivers Point *V* in each drawing indicates the place where the high side of the vertical input of the oscilloscope is generally placed. In some cases manufacturers suggest placing the oscilloscope at the plate of the first video amplifier.

The f-m output cable of the sweep generator is connected to the grid of the video i-f amplifier preceding the last i-f stage to be aligned. The marker generator is also used because it is the only way to tell the response at different frequencies. The exact marker frequencies to be used depend upon the intermediate frequencies of the receivers, and are usually given by the manufacturer in his alignment notes. These frequencies center around 25 Mc, 35 Mc, and 45 Mc. The frequency control of the sweep generator is set to the i.f. of the circuit to be aligned and the sweep-width control advanced to the point where more than the complete curve will be seen. The amount of sweep width used is generally about 10 Mc.

If the video i-f circuits are stagger-tuned, the resonance curve of each circuit will probably be single peaked. If the circuits are all overcoupled then each will have a double-peaked resonance curve. If the marker-generator frequency is the same as that of the resonant frequency of the tuned circuit, then, when the circuit is properly aligned, the marker will be seen at the peak of the curve. In other words, in single-peaked circuits proper alignment will be indicated when the curve has its maximum response at the required resonant frequency of the tuned circuit. The double-peaked curve that results when the *last* stage is overcoupled can be either symmetrical or asymmetrical in shape. For double-peaked asymmetrical curves it is difficult to tell when proper alignment is attained unless we know the exact shape to look for and where certain frequencies are supposed to fall. Examination of actual single- and double-peaked alignment curves of different television receivers will indicate the variations that exist.

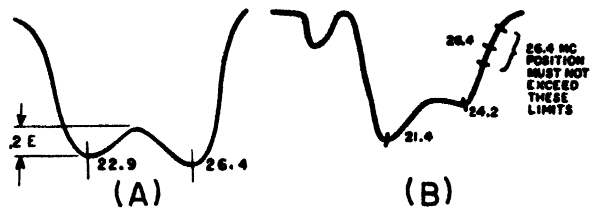


(A) Courtesy RCA (B) Courtesy General Electric (C) Courtesy Capehart-Farnsworth

Fig 16-72.—Response curves of the last video i-f transformers in three different stagger-tuned television receivers.

The three curves of Fig. 16-72 all represent the response of the last video i-f transformer but each is from a different stagger-tuned television receiver. Note how the exact required shape varies even though each is *single peaked.* The marker at the peak of each curve represents the resonant frequency of the tuned circuit in question. Curves (A) and (C) were obtained from conventional type receivers. The marker at the left side of each curve represents the effect of proper tuning of the adjacent video i-f trap. The attenuation afforded by these traps should be such that there is little or no response at these frequencies, as indicated in the drawings. The curve of part (B) is from an intercarrier type receiver. It can be seen from these drawings that it is easy to align such transformers without knowing the exact shape of the curves. All that need be known is the resonant frequency of the video i-f stage and the frequencies of any traps that are employed in this stage.

The two curves of Fig. 16-73 represent the double-peaked responses of the last video i-f stages of two different receivers. Curve (A) is a symmetrical double-peaked response whereas curve (B) is an asymmetrical response. For correct alignment of these stages not only should the i.f. stage of the proper receiver have the shape shown but the markers should fall as indicated. An interesting feature of curve (A) is the indication of the manufacturer that the voltage difference between the peaks and trough should be about 0.2 of the maximum amplitude of the curve. For curve (B)



(A) Courtesy Crosley Corp (B) Courtesy Du Mont Labs.

Fig. 16-73.—Double-peaked response curves of the last video i-f stages of two different receivers

the 26.4 Mc marker (which happens to be the video i-f carrier of the receiver) can fall within the specified set of limits. The small peak on the left-hand side of curve (B) tells us that traps are employed between the last video i-f stage and the video detector. Curve (A) does not indicate such traps even though both curves are taken from conventional type receivers.

Other Video I-F Stages

After the last i-f stage is aligned, the only change made in the connections is the placement of the f-m output cable at the grid input to the video i-f amplifier stage preceding the one to which it was previously connected. The frequencies of the sweep and marker generators are changed to those required for the alignment of the next stage. In visually aligning this stage, whether it be loosely coupled or overcoupled, it should be remembered that the final response curve is the result of the combination of the last two video i-f stages. This means that the response seen on the oscilloscope will be different from that of the stage by itself. It is difficult at this point in the alignment procedure and also for the rest of the alignment (excluding the final over-all video i-f curve) to discuss response shaping that may fit all receivers. The shape of this curve should be made to conform to that required by the manufacturer.

After this stage is aligned, the f-m output cable of the sweep generator is moved to the stage preceding the one just aligned and the necessary frequency changes made in the generators. This method is continued until the very first video i-f transformer is to be aligned. In each case the alignment is made to produce the recommended response. Some manufacturers suggest shorting the video i-f coil preceding the one to be aligned (by simply placing a jumper across it) so that it will not have any possible influence on the response curve. The short is removed when that stage itself is to be aligned. This procedure should be carried out if it is found that the tuned stages preceding the one being aligned cause trouble.

After all these alignments are made, the first video i-f transformer, often referred to as the converter transformer, is aligned last. For this stage the f-m output cable of the sweep generator is connected to the grid circuit of the converter tube. In some receivers it may be very difficult to get the cable at the grid pin to this stage because the converter tube is usually located in the front-end tuner, which is a mechanically compact unit. One method of getting around this difficulty is to make a simple adapter. Take a piece of shielded wire about 3 inches long, strip off about one-half inch of insulation, tin the exposed wire and bend it in the form of a small loop and solder the loop as indicated in the exaggerated drawing of Fig. 16-74A. The actual mean diameter of the loop should be about 1/16 of an inch. The loop is then inserted under the grid pin of the mixer tube and the rest of the wire held against the tube with a rubber band or a piece of scotch tape. The tube is then inserted in its socket. The f-m output cable of the sweep generator is connected or coupled to this wire.

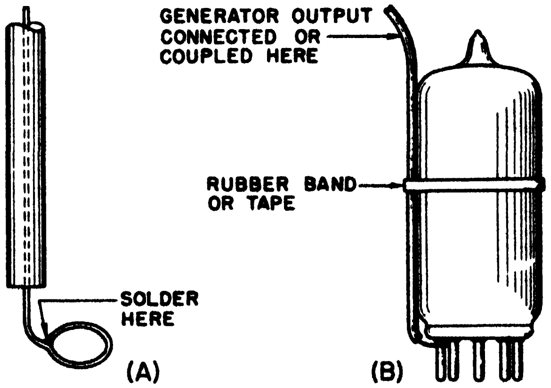
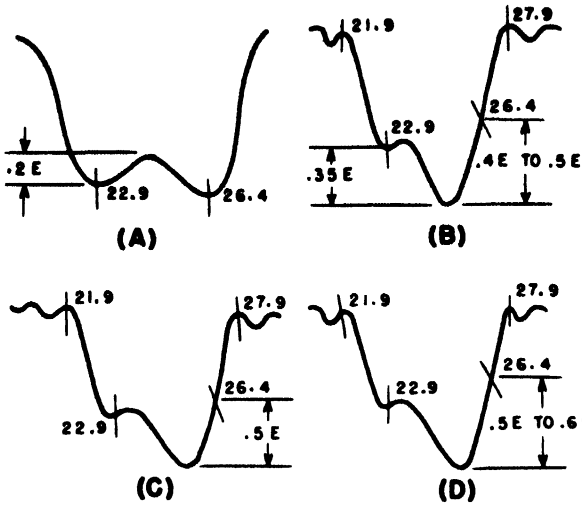


Fig. 16-74. — Simple adapter for aligning the converter transformer.

After proper connections have been made, the first video i-f transformer is aligned to produce a final overall video i-f response curve that resembles the ideal curve of Fig. 16-70A as closely as possible. There are quite a number of acceptable variations in this final response curve, as we shall soon see. Before we discuss these variations let us study a series of typical video i-f alignment pictures recommended by different manufacturers for their receivers.

The four curves of Fig. 16-75 are for a receiver employing overcoupled video i-f stages. Four video i-f transformers and three video i-f amplifiers are used.



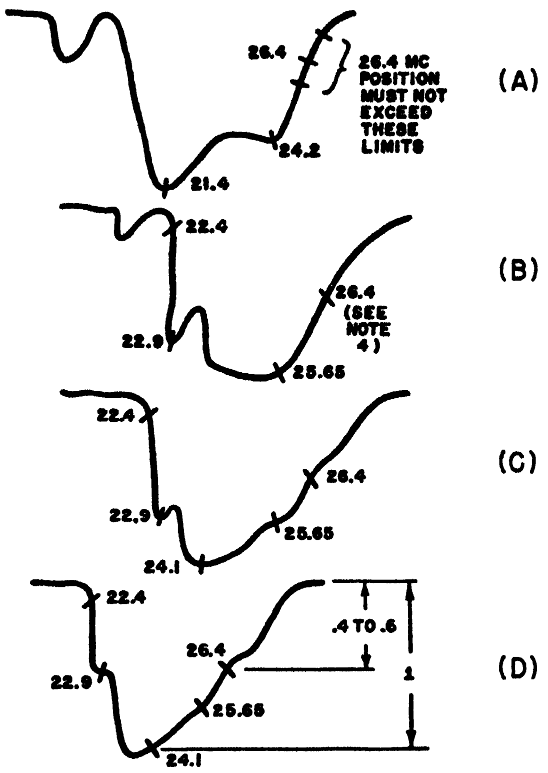
Courtesy Crosley Corp.

Fig. 16-75.—Response curves taken at various points in the video i-f section of a conventional receiver employing overcoupled video i-f stages.

Curve (A) is the response of the last i-f stage, curve (B) that of the last two stages, curve (C) is for the second, third, and fourth stage, and curve (D) is the over-all video i-f response. Note the suggested amplitude characteristics and required marker positions. For comparison purposes we have illustrated in Fig. 16-76 four video i-f alignment curves of another receiver also employing four overcoupled stages. They are illustrated in the order, (A) through (D), corresponding to the previous drawing. Note how the shapes of the curves for the same stage in each drawing differ from each other.

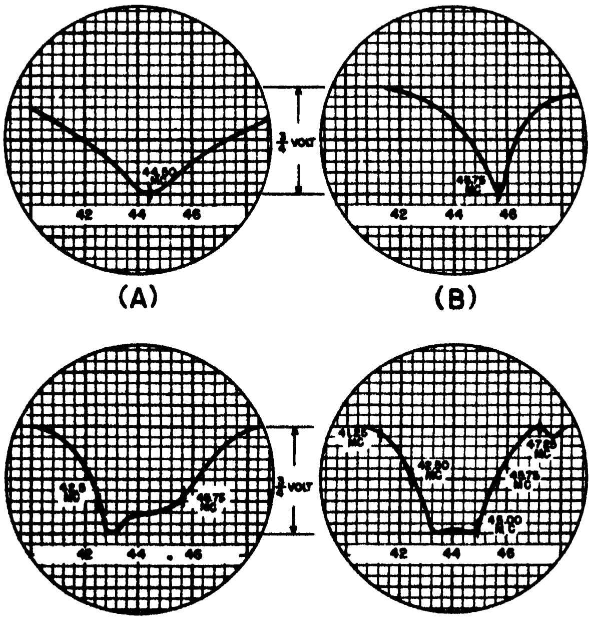
The curves of Figs. 16-75 and 16-76 are for overcoupled video i-f stages of conventional-type receivers. The response curves of Fig. 16-77 are those required for a typical intercarrier television receiver employing stagger-tuned circuits. The same number of stages are employed in this receiver as in the previous one, and the curves are in the same order as in the previous figures. Notice that a 47.25-Mc marker is indicated in the over-all response of curve (D). The dip in the curve at this frequency is due to an adjacent sound i-f trap which is employed between the converter and first video i-f stage of this intercarrier receiver.

Comparison of Figs. 16-75, 16-76, and 16-77 illustrates the marked differences in the shapes of the responses required for video i-f alignment of different television receivers. Although the curves are shown inverted in the figures they may appear upright in practice. Whatever the position of the curve, its frequencies can very easily be determined by the use of marker signals.



Courtesy Du Mont Labs.

Fig. 16-76.—Various i-f response curves of a different conventional television receiver, employing overcoupled i.f.'s, are shown here for comparison with Fig. 16-75.



Courtesy General Electric

Fig. 16-77.—Video i-f response curves of an intercarrier-type receiver employing staggered tuning.

Over-All Video I-F Alignment

Part (D) of Figs. 16-75 through 16-77 illustrates over-all video i-f response curves. Although they all appear markedly different in shape, each one nevertheless produces the desired results. Let us analyze these three curves plus the three new over-all video i-f curves of Fig. 16-78 and see how each conforms to its requirements. While we are doing this, let us bear in mind that the frequency response is from left to right whether the curve is upright or inverted.

Curves (A) and (B) of Fig. 16-78 and (D) of Figs. 16-75 and 16-76 are for the conventional-type receivers whereas curve (D) of Fig. 16-77 and curve (C) of Fig. 16-78 are for intercarrier receivers. On the basis of these curves we can summarize the important requirements and comments concerning over-all video i-f response.

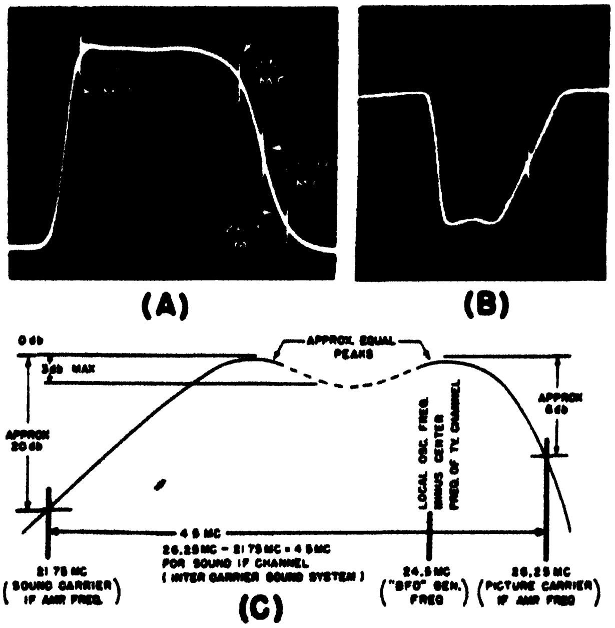
1. The video carrier has to fall on the right slope of the curve somewhere between 40 and 60 per cent of the average maximum response. It is preferable to have this response as close to 50 per cent as possible. A marker signal at the video i.f. is used to check this point on the curve.

2. For conventional-type receivers there should be no response at the accompanying and adjacent sound i-f carrier and adjacent video i-f carrier frequencies. This can be checked by marker signals at the proper frequencies. Even though the i-f traps may have been previously aligned, they may require a slight retuning to bring the response curve to the desired shape. The effects of these traps are generally indicated on the extremes of the curves by small dips.

3. The shape of the response about the midfrequencies of the curve is not too critical, as can be seen from the various curves illustrated. Ideally, the response at these frequencies should be flat but variations are allowed as indicated. The response at the midfrequency, however, should not dip too low.

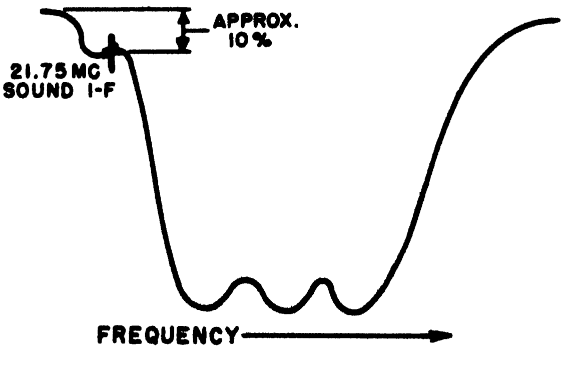
4. For intercarrier receivers there should be a certain amount of response at the sound i-f carrier frequency. This is evident from Figs. 16-77D and 16-78C. The response of both curves at the sound i-f is approximately 10 per cent of maximum. It will be noted that the left side of each curve does not contain a step but has a gradually decreasing slope. An over-all video i-f response curve of an intercarrier television receiver that uses a *step* on the left side of the curve, whose amplitude is at 10 per cent of the maximum response, is shown in Fig. 16-79. A 21.75-Mc marker is shown spotting the sound i.f. of the receiver.

5. Wherever it is deemed necessary, marker signals should be employed to check frequency points along the curve other than those of the accompanying and adjacent sound and video i-f carriers. For example in the curve of Fig. 16-75D, a 22.9-Mc marker is used to indicate the frequency at which the smaller peak should fall. In Fig. 16-76D four extra marker signals, at frequencies of 22.4, 22.9, 24.1, and 25.65 Mc, are used to indicate other parts of the response curve. Extra marker signals are also evident in Figs. 16-77D and 16-78A.



(A) Courtesy RCA
(B) Courtesy Stromberg-Carlson Co.
(C) Courtesy The Hallicrafters Co.

Fig. 16-78.—Three additional over-all video i-f response curves; (A) and (B) are for conventional-type receivers, and (C) is for an intercarrier-type receiver.



Courtesy Capehart-Farnsworth

Fig. 16-79.—Over-all video i-f response curve of an intercarrier receiver that uses a *step* on the left side of the curve.

Intercarrier Receivers Tuned Above and Below the Incoming Signal

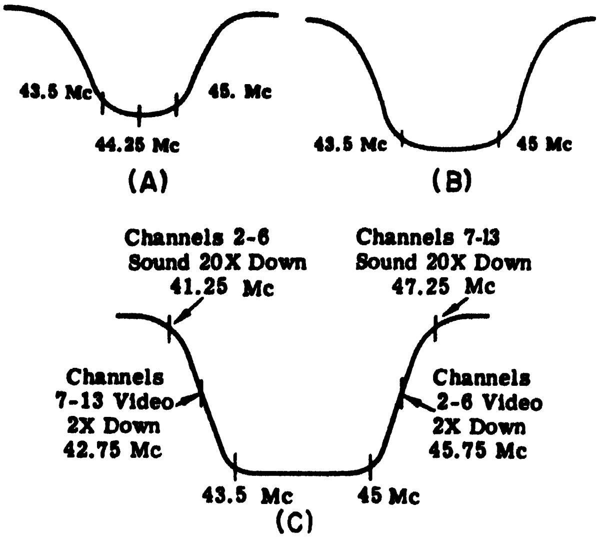
In Fig. 16-70C we illustrated the ideal over-all video i-f response curve for intercarrier receivers where the local oscillator is tuned above and below the incoming signal for the low and high bands, respectively. The alignment of the video i-f stages to produce such over-all curves is done in a manner similar to that previously discussed. The same type of prealignment adjustments have to be made. The connections and control manipulations of the sweep generator and oscilloscope are also the same.

The exact manner in which the curve shapes up as each stage is aligned will vary according to the type of coupling employed. Whatever the case may be, in the end the over-all video i-f response should be symmetrical about the midfrequency of the curve. Let us examine a series of video i-f curves of a typical intercarrier receiver employing such tuning. The patterns appear in Fig. 16-80.

Curve (A) is the response of just the last i-f transformer. Note the required symmetry and positions of the markers. Curve (B) is the resultant response of the last three i-f transformers of the receiver. Notice that the required symmetry is still carried through. The final over-all video i-f response is illustrated by curve (C). This response includes four video i-f amplifier stages. The use of marker signals is very important in checking these curves. The sound and video i-f carriers for the lower channels are 41.25 Mc and 45.75 Mc, respectively; the sound and video i-f carriers for the upper channels are 47.25 Mc and 42.75 Mc respectively.

Note that the video i-f carriers are required to fall "two times down" (2X) which means that their amplitude should be about 50 per cent of the maximum amplitude. Also note that both sound i-f carriers are marked as having a response at 20 times down (20X) which is five per cent of the maximum amplitude. The 43.5-Mc and 45-Mc markers are used to identify those parts of the curve where the amplitude starts to decrease from its maximum value.

Although the over-all video i-f response indicated above is for a particular receiver it can represent the type of curve that should be sought in those intercarrier receivers where response curves are not available. In the alignment of such receivers, the most important criterion is the shaping of the sides of the curve; this shaping should be such that the i-f markers fall as indicated in Fig. 16-80C. Slight variations of these marker points are permissible but the curve should remain as symmetrical as possible. The response between the two slopes of the curve should ideally have a level characteristic, as indicated between the 43.5-Mc and 45-Mc markers in Fig. 16-80C, but variations can exist. There may be a dip in the center of the curve but its height should not be less than 70 per cent of the maximum amplitude.


43.5 Mc
45. Mc
44.25 Mc
(A)
43.5 Mc
45 Mc
(B)
Channels 2-6 Sound 20X Down 41.25 Mc
Channels 7-13 Sound 20X Down 47.25 Mc
Channels 7-13 Video 2X Down 42.75 Mc
Channels 2-6 Video 2X Down 45.75 Mc
43.5 Mc
45 Mc
(C)

Courtesy Zenith Radio Corp.

Fig. 16-80. — Illustrating the development of the over-all video i-f response curve of an intercarrier receiver tuned above and below the incoming signal.

Single-Stage Video I-F Response

It may often be desirable to align the video i-f system by visually observing the response of each i-f stage by itself. There are two methods by which this can be accomplished. One method is to use a traveling probe detector and align each stage individually for the required response. For this method the f-m output cable of the sweep generator is connected to the grid of the video i-f amplifier preceding the i-f tuned circuit to be aligned. The receiver side of the probe detector is connected to the plate circuit of the video i-f amplifier following the i-f tuned circuit being aligned. During this process marker signals can be used to identify frequency points on the curves.

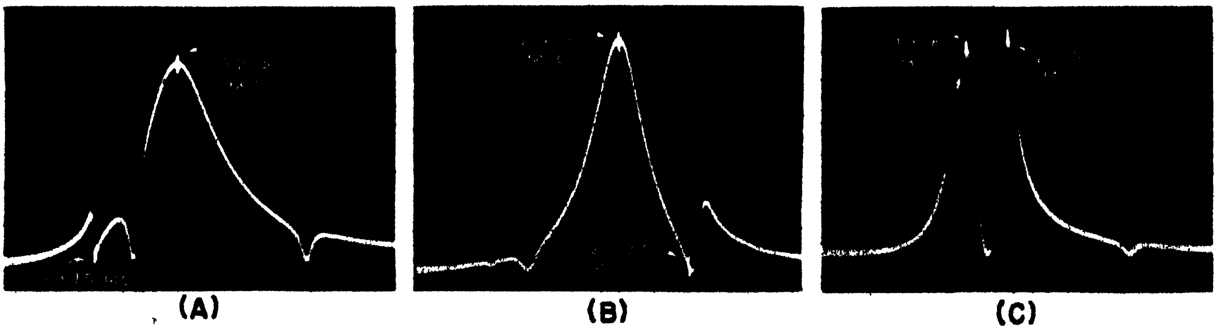
The other method uses the video detector of the receiver instead of a probe detector. The tuned circuits in the alignment path, other than the one being adjusted, are loaded down to the point where they have negligible response characteristics. There are two possible alignment procedures for this latter case. One is to connect the f-m output cable of the sweep generator to the converter grid circuit and this connection remains as is for the complete video i-f alignment procedure. The vertical terminals of the oscilloscope are placed across the video detector load and the modulating sine-wave output from the sweep generator fed to the horizontal input of the oscilloscope. The video i-f transformers and tuned circuits not being aligned are shunted with resistors of 300 to 500 ohms to load them down.

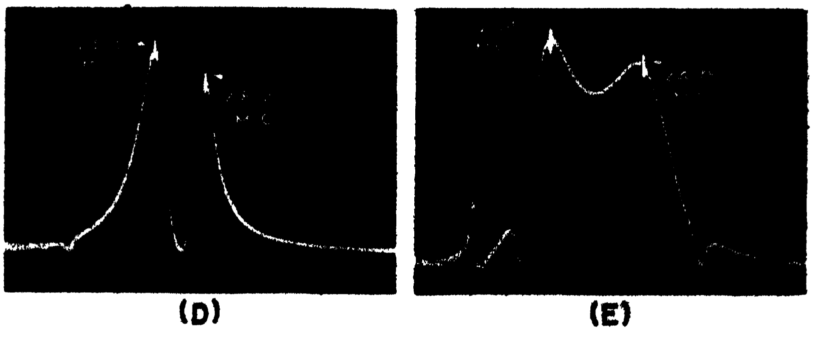
The frequency control and sweep-width control of the sweep generator are adjusted to produce a complete pattern upon the oscilloscope. With the desired pattern known, the proper video i-f circuit is aligned to produce this pattern; markers should be used to check frequency points. After one stage is aligned it is shunted with a resistor and the shunting resistor across the next stage to be aligned is removed. The frequency and sweep-width controls of the sweep generator are readjusted, if necessary, to produce a complete curve on the oscilloscope. The circuit is then aligned to produce the required response. This method of alignment is carried out for each video i-f stage. The sequence of alignment is not too important but, for reasons of simplicity, it is suggested to start from the last i-f transformer and work backward.

For the other procedure the sweep generator is placed at the grid input to the stage preceding the video i-f circuit to be aligned. Only those i-f tuned circuits other than the one being adjusted, which are in the path of the input signal are loaded down. After each stage is aligned, it is loaded down with a low-valued resistor and the sweep generator is moved to the amplifier grid preceding the i-f transformer to be next aligned.

Once the alignment of each i-f stage is carried out in this manner, all the loading resistors are removed and with the sweep generator remaining as in the last alignment, that is at the converter grid, the over-all video i-f response is checked. This over-all response should then appear as the required resultant waveshape. A small amount of tuned circuit readjustments may be necessary to reshape the curve slightly so that its response around certain marker frequencies will be correct.

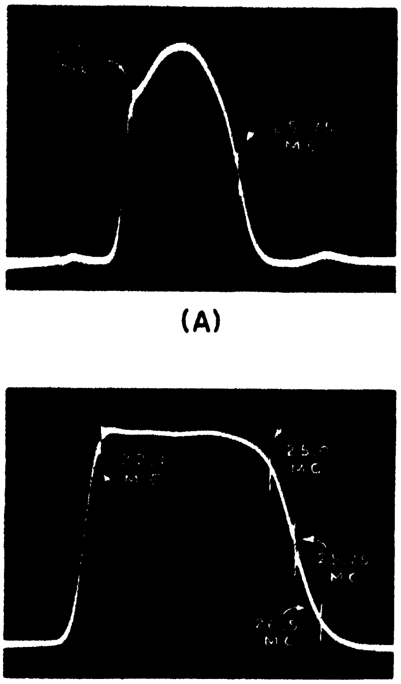
As an example of how such individual alignment curves appear, we have included some actual pictures from a typical stagger-tuned conventional-type television receiver. These are in Fig. 16-81. They are arranged starting with the response of the last i-f transformer and continuing backward. Curve (A) is the response of the fifth video i-f transformer, which is the last one of the receiver. Curves (B), (C), and (D) are the responses of the fourth, third, and second video i-f transformers, respectively. The video i-f transformer arrangement lying between the converter and first video i-f amplifier is somewhat different from the others and possesses a double-humped characteristic as indicated by pattern (E). Note the marker signals pointing out where certain frequencies should appear. The large dips in the center of the curves of patterns (C) and (D) are the results of sound i-f traps em-


(A) (B) (C)


(D) (E)

Courtesy RCA

Fig. 16-81.—Video i-f response curves of the individual stages of a typical stagger-tuned conventional-type television receiver.



(A)

(B)

Courtesy RCA

Fig. 16-82.—Other video i-f response curves of the receiver of Fig. 16-81: (A) is for all of video i-f stages except the first; (B) is the over-all video i-f response.

ployed in these stages. A 21.25-Mc accompanying sound i-f trap causes the dip in pattern (C) and a 27.25-Mc adjacent sound i-f trap causes the dip in pattern (D).

For comparison purposes we have included the over-all response of all these i-f transformers except the first (i.e. the second, third, fourth, and fifth) in part (A) of Fig. 16-82. In part (B) of Fig. 16-82 is the complete over-all video i-f response curve of all the i-f stages. We can consider this latter curve also as being the combination of the curves of Figs. 16-81E and 16-82A.

R-F Alignment

The type of tuner employed in a television receiver determines which shall be aligned first, the r-f tuned circuits or the oscillator. To pin down exactly which comes first for different types of tuners is a difficult task. We are going to follow the procedure recommended by the majority of television receiver manufacturers, namely, the r-f alignment first and the oscillator alignment last. In some receivers r-f and oscillator alignment is not required because the tuner is employed in a factory-sealed unit. Seldom will the r-f and oscillator stages require alignment. However, if alignment of these sections is found to be necessary, the receiver should be sent back to the factory or to an authorized dealer.

The most important thing in r-f alignment is to make sure that the tuned circuits for each channel are broad enough and fairly level to be adequately responsive to both the video and sound r-f carriers. For any particular channel these carrier frequencies are the same no matter what receiver we are dealing with. These frequencies are determined by the transmitter. The video and sound carriers are always 4.5 Mc apart, with the sound carrier always higher than the video carrier by this amount. In Table II appear the different television channels in use and the assigned frequency limits of each channel. In the third and fourth columns are listed the transmitted video and sound-carrier frequencies, respectively. During the process of visually aligning the r-f stages for any one channel, a marker generator is required with signal frequencies equal to the video and sound r-f carriers of this channel. This is necessary in order to observe the response at these frequencies.

TABLE II

CHANNEL NO.	CHANNEL FREQUENCY (MC)	VIDEO CARRIER FREQUENCY (MC)	SOUND CARRIER FREQUENCY (MC)
2	54—60	55.25	59.75
3	60—66	61.25	65.75
4	66—72	67.25	71.75
5	76—82	77.25	81.75
6	82—88	83.25	87.75
7	174—180	175.25	179.75
8	180—186	181.25	185.75
9	186—192	187.25	191.75
10	192—198	193.25	197.75
11	198—204	199.25	203.75
12	204—210	205.25	209.75
13	210—216	211.25	215.75

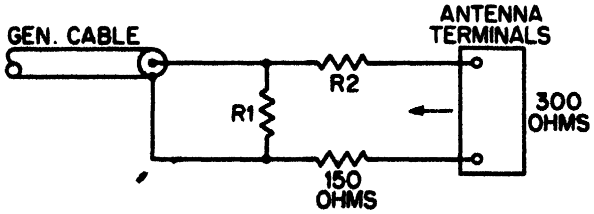
Prealignment Adjustments and Connections

The alignment of the r-f section of the receiver has its share of prealignment adjustments as well as the other sections. The agc voltage on the r-f tubes, if any such voltage is used, should be replaced by either a bias battery or from elsewhere within the receiver. The manufacturer usually recommends the best method to use on his receiver. The contrast control should be set so that it does not cause overloading of any of the tubes in the receiver.

In most r-f alignment procedures, the f-m signal from the sweep generator is fed to the antenna input terminals of the receiver. The method of connection is very important although not all manufacturers stipulate this. For proper alignment the output impedance of the sweep generator and the input impedance of the receiver should be the same, that is, they should be matched. However, since the input impedances of television receivers vary, so will the output impedance of the various sweep generators be different. Whatever the case may be, it is possible to properly connect the sweep generator to the antenna input terminals through a special matching network.

We have to consider whether or not the sweep generator has provision for balanced and/or single-ended output and whether the input to the receiver is balanced or not; in the majority of cases, the receiver input is balanced. If the output impedance of the sweep generator is balanced and equal to that of the balanced receiver input, the sweep-generator cable is connected directly to the antenna input terminals, that is, no matching network is needed. There are some receivers that have provision for a choice of two input impedances, one usually being balanced and the other unbalanced. The unbalanced input is usually of lower impedance and sweep generators with single-ended outputs can be used with this type of receiver input.

If the sweep generator used has a single-ended output, it can be matched to a balanced receiver input in the manner shown in the circuit of Fig. 16-83. For this drawing the receiver input is assumed to be 300 ohms (as it is in the majority of cases). Three resistors are employed in the match, with one resistor inserted across the sweep-generator output cable. One of the resistors is made equal to 150 ohms and the other two are calculated on the basis of the output impedance of the generator. It is usually the cable employed with the generator (which is normally of the coaxial type) that determines the output impedance. In some cases, resistor *R1* may be employed within the sweep generator.



GEN.	R1	R2
50	56	120
72	82	110
92	110	100

After General Electric

Fig. 16-83.—Circuit and table of values for matching an unbalanced sweep generator output to a 300-ohm balanced receiver input.

The table in Fig. 16-83 indicates the desired values of resistors *R1* and *R2* for different values of generator output impedance, for a receiver with a balanced input impedance of 300 ohms. The balance can be checked by calculating the impedance seen looking in the direction of the arrow, from the antenna terminals to the generator. When looking in this direction, the total impedance seen is equal to the parallel impedance of *R1* and the generator, plus the series impedance of *R2* and 150 ohms.

As an example let us assume the generator impedance to be equal to 50 ohms, hence *R1* and *R2,* as seen from the table, become equal to 56 ohms and 120 ohms, respectively. The parallel impedance of 50 and 56 ohms is approximately 29 ohms and when added to the series impedance of 120 ohms and 150 ohms, the total impedance becomes 299 ohms. Thus we see how such a match is obtained with a 300-ohm balanced input receiver. The values given in the table are not too critical and slight variations are permissible. The primary requisite in any match of this type is that the impedance seen in the direction of the arrow approach that of the receiver input impedance.

The frequency adjustment of the sweep generator should be set to the center frequency of the channel being aligned. This center frequency can be very easily calculated from the second column of Table II. For example, for channel 2, the center frequency is 57 Mc; for channel 3, it is 63 Mc; for channel 11, it is 201 Mc and so on. The sweep of the generator should be capable of sweeping over a complete channel. Since it is advisable to see more than the complete curve, a sweep width of about 15 Mc is suggested. If the generator is not capable of such a sweep width, the maximum sweep available should be used. The amplitude control of the f-m output signal is adjusted to produce a noticeable deflection on the oscilloscope without causing any overload of the receiver stages. The sine-wave modulating signal output from the generator is connected to the horizontal input terminals of the oscilloscope.

Unless otherwise noted in the manufacturer's service notes, the high-frequency oscillator is made inoperative either by pulling out the oscillator tube, grounding the oscillator grid, or by any other method. Whatever method is used, it should not cause any impedance changes or loading effects on the tuned r-f circuits. The fine tuning control should be set at the center of its range. There are unusual cases where the precision alignment of a single oscillator stage (usually channel 13) is required before alignment of the r-f stages is started.

Connecting the Oscilloscope

In the alignment of the r-f stages there are three principal methods of connecting the oscilloscope. In each case the vertical terminals of the oscilloscope must follow a stage of rectification. One common method is to connect the vertical terminals of the oscilloscope across the video detector load. The response curve appearing on the oscilloscope for each r-f alignment in this case will be the resultant of the over-all video i-f response plus that of the r-f response. For this procedure the video i-f stages must be correctly aligned.

Another method makes use of a probe detector. The receiver terminals of this detector are usually connected to the output circuit of the converter tube. In this manner only the response of the r-f tuned circuits will appear upon the screen of the oscilloscope. If the probe is to be used in any other r-f section of the receiver, care must be taken that it does not cause any loading effects on the r-f tuned circuits.

A third method makes use of grid-current flow in the converter circuit during the alignment process. In effect the cathode and grid of the converter act as a rectifier in a manner similar to the action of the grid and cathode of a limiter tube. In a situation such as this, the vertical terminals are usually placed across the proper grid resistor in the converter circuit. If the grid resistor is not of the proper value or if it is not easily accessible, many manufacturers insert a special resistor or testing point for this alignment process. If the grid resistor used in the circuit is low, then insert a large resistor (about 100,000 ohms) in series with the high side of the vertical input terminals. This resistor is used to prevent the oscilloscope from loading down the grid circuit of this tube. Sometimes the oscillator section of the receiver is made operative because in certain cases, only with the oscillator working, is it possible to have grid current flow in the converter circuit suitable for r-f alignment purposes.

Sequence of Alignment

In many cases the alignment for the tuned r-f stage of one channel will affect the alignment of other channels. This is so because, in many instances, the inductances of all or some of the r-f tuned circuits are part of a single coil, where the desired amount of inductance is chosen by a switch or a continuous tuner. A simplified drawing of a case where a switch is used is illustrated in Fig. 16-84. The tuning capacitors are omitted from this drawing. The coil in the grid of the r-f tube represents the tuned antenna stage and the other coil is part of the second tuned r-f stage. The coils shown are for the high television bands only (channels *7-13*); a similar arrangement is assumed for the low band.

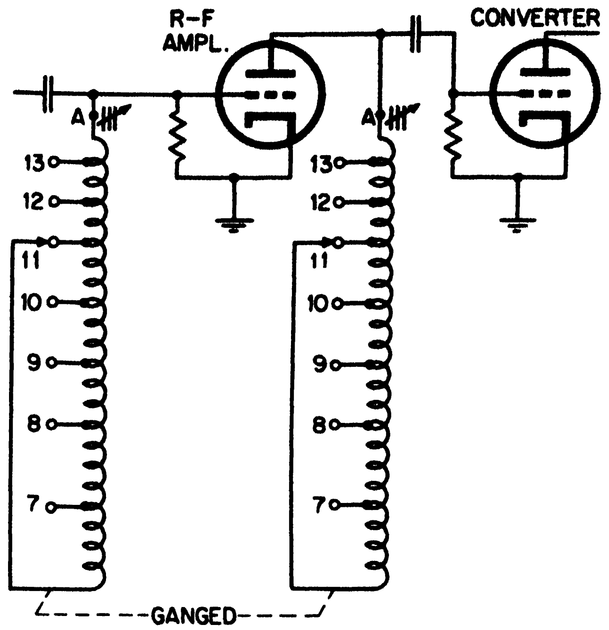


Fig. 16-84—R-f stages in which channel tuning is accomplished by tapping a single coil.

Since channel *13* has the highest frequency, the inductance of its tuned circuit will be the smallest and hence channel *13* is normally aligned first. The switch in the drawing of Fig. 16-84 should be connected to terminal number *13* for this operation. After channel *13* alignment, channels *12, 11, 10,* etc., are aligned in descending order. In the alignment of any one channel, the inductances of all the channels *lower* in number than the one being aligned are shorted out by the switch. This means that when the switch is on a particular channel, for example channel *11,* as shown in the drawing, only the inductance between point *A* and tap number *11* on the coil is effective for the r-f tuned circuits of channel *11.*

From this method of tuning we see that the inductance of any tuned circuit includes the inductances of the channels higher in frequency than the one being aligned. Consequently we see that the alignment of any one channel depends upon the alignment of the higher channels. In the drawing, with the tuning switch set to channel *11* for alignment purposes, the r-f tuned circuits of channels *13* and *12* should already have been aligned.

In many television receivers only one or two inductances may be varied for a complete set of channels (high or low). In a situation such as this, it is equally important to align the proper channel first. The channel aligned first is the one that has its inductance variable and is usually the higher-frequency channel of the band in question. It should also be remembered that this variable inductance may be part of the tuned r-f circuit of other channels.

Alignment Procedure

In the alignment of the r-f section of a television receiver, as well as the oscillator section, it is very important to have a diagram of the trimmer locations. The word "trimmer" as it is used here refers to both capacitive and inductive tuning components. Such information is usually given in the service notes of the receiver. With the trimmer locations known, the general r-f alignment procedure is as follows.

With the sweep generator properly connected to the antenna input terminals, set the generator frequency to the center of the first channel to be aligned. As mentioned previously, this channel is usually the one that has its tuned circuit inductance variable. The oscilloscope may be placed across the video detector load, at the converter grid as outlined previously, or a detector probe may be used.

If the oscilloscope is placed at the grid of the converter tube, marker signals of the exact video and sound r-f carriers of the channel being aligned (see Table II) should be fed into the receiver. The proper coils and capacitors are then tuned to produce a curve on the oscilloscope that approximates the ideal response shown in Fig. 16-85. If there is no recommended order of adjusting these trimmers, then it is suggested that the necessary coils and capacitors of the last r-f stage be aligned first, and then, working backward, the other r-f stages (for the same channel) are aligned. The pattern shown is that required by many manufacturers. Such a pattern indicates that the r-f stages are overcoupled. The trough of the curve should have an amplitude no smaller than about 70 per cent of the maximum amplitude of the curve. The capacitor trimmers are usually tuned to make sure the curve is positioned properly about the markers and also to help produce the proper bandwidth. The coils, in conjunction with the capacitors, are adjusted to produce as good a flat top characteristic as possible.

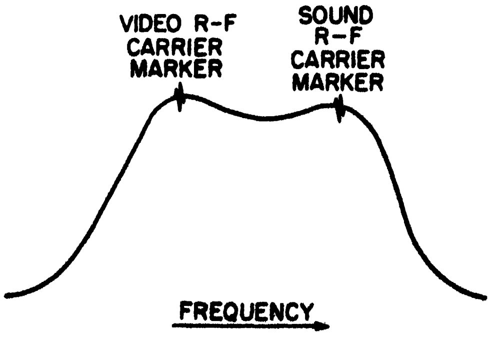


Fig. 16-85.—Ideal television r-f response curve.

In some receivers this double-peaked response is required only for the low-band channels and a different response, usually a broad single-peaked curve, is required for the high-band channels. The exact position of the two markers will tell us how well the r-f circuits are tuned. If both of these markers fall within the two peaks of an overcoupled curve or if the response of the curve at these frequencies is above a certain predetermined value, then the alignment is assumed to be correct. Some manufacturers specify that the response at the marker frequency for the *first* channel aligned in each band should be no less than 90 per cent of maximum. This limit helps insure a proper response for the other channels.

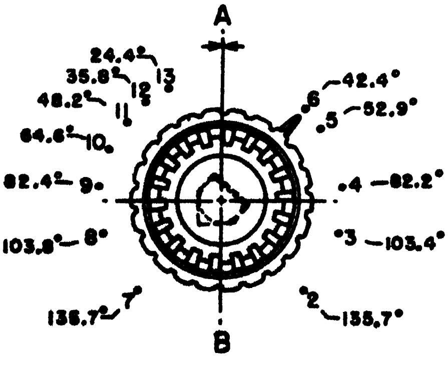
Once the first channel, for either the high or low band, is properly aligned, the frequency control of the sweep generator is set to the center frequency of the next channel to be aligned and the receiver tuned to that channel. The sound and video r-f marker signals are changed to those of the new channel to be aligned. If a fine tuning control is incorporated, set it to the center of its range. If the receiver has a switch type of tuner, it is a simple matter to select the proper channel. However, if the channel selector is of the continuously variable type, it may be somewhat difficult to find the exact point where the selector is set to the center of the channel. Some manufacturers realize this and include a diagram of the front section of their tuner indicating at what points in the rotation of the selector knob the channel center frequencies will be found. A typical diagram of this sort appears in Fig. 16-86.

The channel locations are indicated by dots around the circumference of the knob, with the proper channel number indicated next to each dot. The exact position of each channel is marked off as being so many degrees away from a theoretical vertical line *A-B* passing through the diameter of the knob. For the example under discussion, the degree measurements are taken with point *A* as a reference. Thus for the low channels we turn the knob clockwise from point *A* starting with channel 6, which is 42.4° away, and end with channel 2, 135.7° away. For the high bands, rotation is counterclockwise from point *A* starting with channel 13, which is 24.4° away, and ending with channel 7, 135.7° away.

With all the proper settings made, a response curve should appear on the oscilloscope. The tuning coils and capacitors of the first stage aligned are usually common parts of all the tuned r-f circuits in each band. This means that when the r-f response of the second channel is checked and it appears correct with the markers falling within limits, then no adjustments are needed for this channel. The responses of the remaining channels in each band are checked in the same manner. If any response falls outside of the pattern recommended, then the adjustable coils and/or capacitors may require a *slight* readjustment. If such a condition is found necessary, then, after the responses of all the remaining channels are examined, the responses of the previous channels should be rechecked to see if they still conform with those required. In each case it is the location of the markers that tells us how accurate the alignment is.

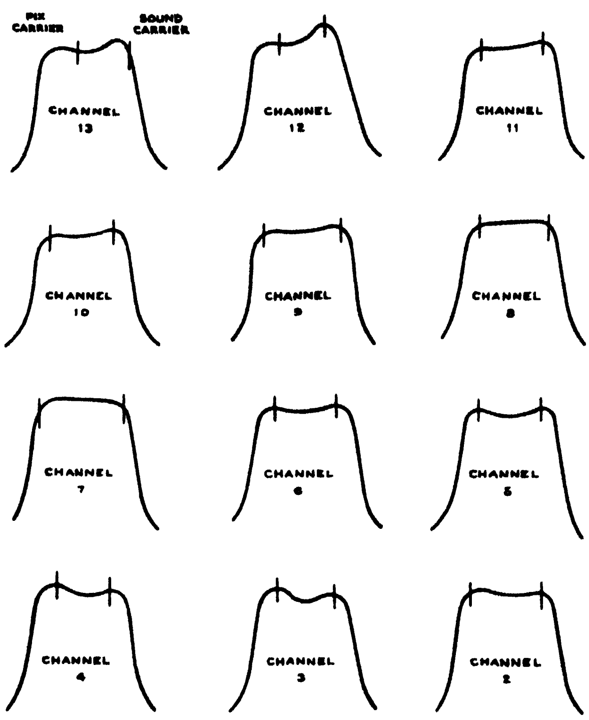
For the upper channels (7 through 13) of most television receivers there is usually only one set of tuning coils and capacitors, to be adjusted during alignment of the highest frequency channel. For the lower channels (2 through 6) the same situation often exists, there is one set of trimmers for the complete low-frequency band. However, there are quite a few receivers where *two* sets of trimmers are used for the low band. One set of trimmers is usually employed for channels 4 through 6 and the other set for channels 2 and 3. Although two separate sets of trimmers are used, the channels are usually aligned in descending numerical order starting with number 6. When the tuned circuits of channels 2 and 3 are aligned, the inductances of the other three channels (4, 5, and 6) are included as parts of these tuned circuits.

The majority of manufacturers show one or two r-f alignment curves to represent all channels. They will also indicate, either by drawings or by statements, the allowable variations in the shape of the response. There are other manufacturers who indicate a response curve for each channel with a statement that the response at the marker frequencies should be above a certain limit for each drawing.



Courtesy Capehart-Farnsworth

Fig. 16-86.—Diagram of a continuously variable tuner knob indicating the angular location of the various channels.

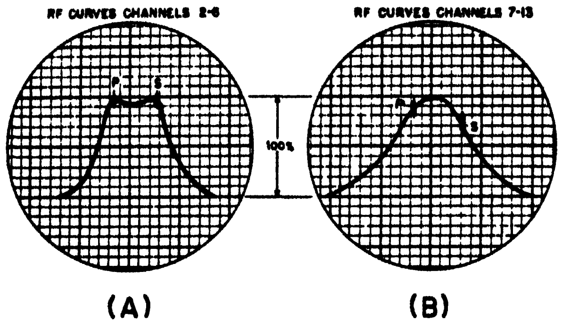


Courtesy RCA

Fig. 16-87.—R-f response curves recommended for each channel of a typical receiver.

In Fig. 16-87 appear a series of r-f response curves, one for each channel, as recommended by one manufacturer. Note how closely the shapes of the curves resemble each other. The small vertical lines on each curve represent the positions of the video and sound r-f markers. The marker at the left of each response curve is that of the video r-f carrier and the marker on the right is that of the sound r-f carrier. The markers of each curve fall well within the limits of the response, indicating that each represents the proper bandwidth. For these curves the manufacturer states that the response at the marker frequencies should not fall below 80 per cent of the maximum response.

The curves in Fig. 16-88 represent a situation where a different type of response curve is required for the high- and low-frequency bands. Both curves represent the ideal shape desired. The letters *P* and *S* in each drawing indicate the video and sound r-f markers. Curve (A) is for the low-frequency channels and curve

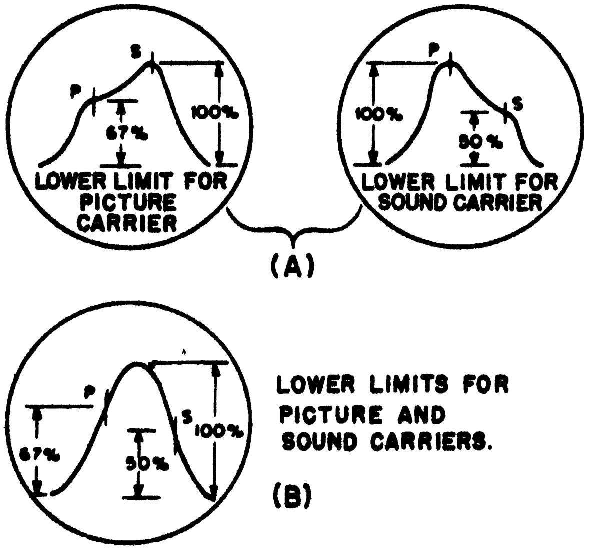


Courtesy General Electric

Fig. 16-88.—R-f response curves recommended for the low-frequency channels (A) and the high-frequency channels (B) of a specific receiver.

(B) for the high-frequency channels. The double-peaked characteristic of curve (A) indicates that the low-frequency channels have overcoupled r-f stages. The single-peaked shape of curve (B) indicates that the r-f stages are loosely coupled for the high-frequency channels. It is possible to obtain the desired 6-Mc r-f bandwidth on the high-frequency channels with single-peaked curves because the required bandwidth is a smaller percentage of the center frequency of these channels, as compared with the low-frequency channels.

The drawings of Fig. 16-89 illustrate the allowable deviations in the shapes of the response curves of Fig. 16-88 and are based upon the positions of the markers. Part (A) is for the low-frequency channels and part (B) for the high-frequency channels. If the shape of the response on any low-frequency channel falls between the ideal curve of Fig. 16-88A and either one of Fig. 16-89A then it is considered acceptable provided the response at the video and sound carriers is above the limits indicated. For the high-frequency channels the drawing of Fig. 16-89B indicates that the response at the sound carrier cannot fall below 50 per cent nor the video carrier fall below 67 per cent of the maximum response.


S
P
67%
100%
LOWER LIMIT FOR PICTURE CARRIER
P
S
100%
50%
LOWER LIMIT FOR SOUND CARRIER
(A)
P
S
67%
50%
100%
LOWER LIMITS FOR PICTURE AND SOUND CARRIERS.
(B)

Courtesy General Electric

Fig. 16-89.—Allowable deviations in the response curves of Fig. 16-88. (A) is for the low-frequency channels; (B) is for the high-frequency channels.

Thus far our discussion concerning r-f alignment was based upon the oscilloscope being placed in the converter grid circuit. If the oscilloscope is placed in the output of the video detector, the alignment procedure is not changed very much. Everything said relative to the sweep generator and marker generator is also true here. For this type of alignment the r-f oscillator has to be operating because heterodyning action inside the converter tube is necessary. The adjustment of the different r-f trimmers follows the same sequence as before. The resultant response will have a different shape than that previously shown. It will appear very much like the over-all video i-f curve. The video and sound *r-f* marker signals, when heterodyned in the converter, come out as video and sound *i-f* marker signals. When these markers are changed to i-f signals, the sound i.f. is *lower* than the video i.f. whereas in their original form, the sound r.f. is *higher* than the video r.f.

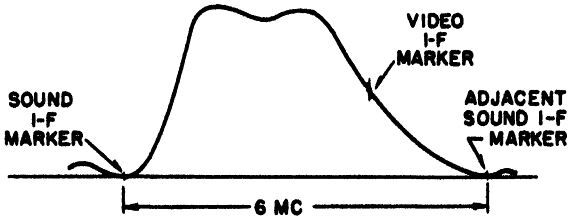


Fig. 16-90.—Ideal over-all r-f and video i-f response characteristics for conventional television receivers.

The ideal over-all r-f and video i-f response characteristics generally recommended for conventional receivers appear in Fig. 16-90. For correct r-f alignment, the video and sound carrier markers should fall within their appropriate limits. Ideally speaking, if the sound i-f marker falls in its proper place, it should be barely, if at all, visible. The video i-f marker should fall on the right slope and have a response anywhere from 40 to 60 per cent of maximum. A check can be made with other marker signals for the location of certain frequencies. One such marker, the adjacent sound i.f., is illustrated in the drawing. This marker, when falling in its proper place, should be barely visible because there should be no response at this frequency.

For intercarrier receivers, with the oscilloscope placed at the video detector, the alignment procedure is similar. The only difference will be in the shape of the final response curve. A typical curve is shown in Fig. 16-91. Note that there is the required response at the sound i.f. The response at this frequency should, in most cases, have a limit of about 10 per cent of the maximum response. The video i-f carrier should fall approximately in the center of the right slope; a 40 to 60 per cent response limitation is allowed here too. The required response is not much different from that generally required for the over-all video i-f section of the intercarrier receiver. If the receiver has a step function at the sound i-f carrier in the video i-f response, then it will appear in a form similar to the over-all r-f and video i-f response curve.

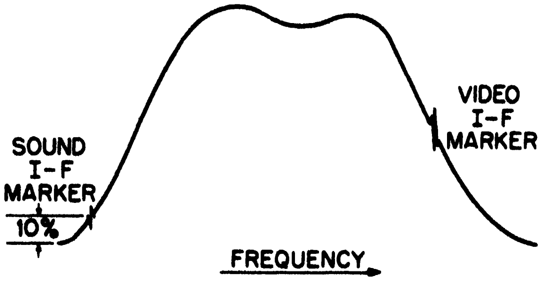


Fig. 16-91.—Ideal over-all r-f and video i-f response characteristics for intercarrier receivers.

If the intercarrier receiver has its oscillator tuned above and below for the lower and upper channels, then the over-all response of the r-f and video stages will very closely approximate that of the over-all video i-f stages alone. This is so because the response at all the sound and video i-f carriers must be the same for both types of curves.

Oscillator Alignment

There are three principal methods that are used today for aligning the oscillators of television receivers. One arrangement, referred to as the visual method, employs an oscilloscope across the video detector load and a sweep signal at the antenna input terminals. A second method employs a vacuum-tube voltmeter or a zero center meter as the indicating device and it is placed across the *sound* detector load. An accurately calibrated r-f generator is used as the source of signal input; a sweep generator is not needed. The third method uses a heterodyne frequency meter only; a sweep or marker generator is not employed.

Although this chapter is on *visual* alignment, the other two methods of oscillator alignment are interesting enough to warrant discussion also. For oscillator alignment it is generally not necessary to make any prealignment adjustments as was necessary for the previous alignments. The only adjustment necessary pertains to the fine tuning control, if any is incorporated. Unless otherwise noted, *this tuning control should be set to the center of its range during the complete alignment of the oscillator section no matter what alignment method is employed.*

If any variable capacitors, other than the fine tuning control, are used in the oscillator section, they are usually in the circuit for every channel. Consequently any one adjustment of such components will generally be sufficient for all channels. In many cases the inductance of the oscillator stage is a single coil for both bands, with taps for switch tuning. This is similar to that discussed for the r-f section. In other cases a separate coil arrangement is used for the high and low bands.

In some television receivers a *single* set of trimmers is used for the high band and another set for the low band; this is similar to some r-f circuits. In alignment of such oscillator circuits, the adjustments are made for one channel of each band and the other channels usually fall in line or require only a slight readjustment of the trimmers. Other receivers have a single set of adjustments for all channels of the high band and a *separate* inductive adjustment for *each* channel of the low band, and other television receivers have a separate inductive adjustment for *every* channel of the receiver. Whatever tuning adjustments are employed, when the oscillator section of each channel is checked, the indicating device should reveal whether or not the oscillator is properly aligned.

Visual Method

In the visual method the sweep generator and marker generator remain as they were for r-f alignment and the oscilloscope is placed across the video detector load. The frequency and sweep width of the sweep generator for each channel is the same as in r-f alignment. The sound and video r-f markers for each channel are also inserted into the circuit. These signals will appear as i-f markers on the response due to the heterodyning process inside the converter tube. The resulting response will be a combined r-f and over-all video i-f response and will take on the shapes previously shown for r-f alignment when the oscilloscope was placed at the output of the video detector.

Proper alignment of the oscillator is indicated primarily by the video i-f marker appearing along the right slope of the curve, at about 50 per cent of maximum, or as otherwise indicated by the manufacturer. The sound i-f marker should be just visible on the left slope for intercarrier type receivers (about 10 per cent of maximum response) and barely visible for conventional receivers. A check on the alignment can be made by turning the fine tuning control a little on either side of its center setting. This should move the video i-f carrier up and down the right or high-frequency side of the response. If it was necessary to readjust some oscillator trimmers that had been previously adjusted, then it is advisable to recheck the complete alignment to see that, for each channel, the proper curve appears with the appropriate markers in their correct positions.

Meter Method

The meter method of alignment to be discussed here is only suitable for conventional-type television receivers and *not* the intercarrier type. In this method of aligning the oscillator section of television receivers, the vacuum-tube voltmeter (or zero center meter) is placed across the load of the sound f-m detector. If the detector uses a discriminator, then the meter is placed across the complete resistive load, across points *A-B* in Fig. 16-60. If it is a ratio detector, the meter is placed between the point of audio output and ground or as indicated in Fig. 16-66. The input to the antenna terminals is an accurately calibrated unmodulated r-f signal whose frequency is equal to the r-f sound carrier of the channel being aligned.

For this type of oscillator alignment the sound section of the television receiver must be accurately aligned first. Proper oscillator alignment is indicated by a zero voltage reading on the meter. What happens is that the oscillator signal heterodynes with the input sound r-f carrier in the converter tube and the difference frequency that results when the oscillator is properly aligned is equal to the operating sound *i.f.* of the receiver. In a *conventional-type* television receiver, the sound i-f transformers are tuned to this difference frequency. Hence when the oscillator for any one channel is tuned to its proper frequency, there will be a zero voltage indication on the meter in the sound-detector circuit.

Heterodyne Frequency Meter Method

For this method of oscillator alignment all that is required is an accurately calibrated heterodyne frequency meter. This is the only method where the knowledge of the exact oscillator frequency for each channel is required; it can be used for *any* type of receiver. The oscillator frequency is not the same for each receiver, as are the sound and video r-f carriers, but differs according to the i.f.'s employed.

For example, if the video i.f. is 25.75 Mc and the oscillator is tuned *above* the i.f. for all channels, then the required oscillator frequency is equal to this i.f. *plus* the *video* r-f carrier frequency for each channel. Therefore, the oscillator frequency for channel 2 should be 25.75 Mc plus 55.25 Mc or 81 Mc; for channel 3, it should be 25.75 Mc plus 61.25 Mc or 87 Mc, and so on. The oscillator frequency can also be found by using the operating sound i.f. of the receiver and adding it to the sound r-f carrier; or if the oscillator is tuned below the incoming signal, subtracting it from the r-f carrier.

The frequency meter is adjusted so that its internal frequency standard is equal to the required oscillator frequency of the channel to be aligned. The fine tuning control, if any is employed, is set at the center of its range. Loose coupling between the frequency meter probe and oscillator circuit is made so that the oscillator signal may be coupled to the meter. The frequency meter used may, however, be sensitive enough to pick up the radiated oscillator signal without making any actual connection between units.

Once these preliminary adjustments are made, the oscillator circuit is aligned to produce a zero beat note in the output of the frequency meter. Since an exact zero beat note may be difficult to obtain, the proper alignment point is indicated by an audible beat note of lowest frequency.

Tube Adaptors in Television Receivers

The tube adaptor is a simple means of attaining contact with tube terminals for voltage and resistance measurements. Although quite old basically — we showed its use about 20 years ago — it is again receiving attention due to the layer form of assembly in some television receivers. For that matter, it is usable in any confined space where access to the socket connections for voltage and resistance measurements is difficult.

As is commonplace, many devices intended for one use find other applications and we can readily conceive of such adaptors being employed in connection with alignment operations in television video i-f circuits, especially when access to the different tubes in the system is not readily available. Since the use of the adaptor does not *prevent* alignment — and the effects of the invisible components introduced by the adaptor become evident only by comparison, we feel that a warning should be issued concerning the use of *any* adaptor in the video i-f systems of television receivers during alignment operations.

Alignment of video i-f systems should be accomplished only with the circuit components exactly as they will be used; anything added to the system will introduce a change between the response curve with it in the circuit and after it has been removed. This is illustrated in Figs. 16-92 through 16-96. These are the over-all video i-f response curves when a very simple adaptor was used in different sockets of the amplifier.

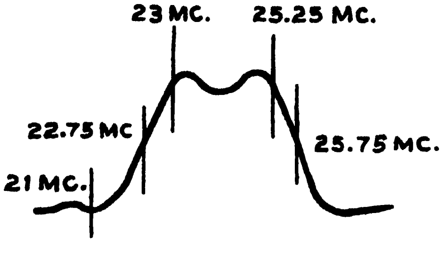


Fig. 16-92.—Over-all video i-f curve without adaptor.

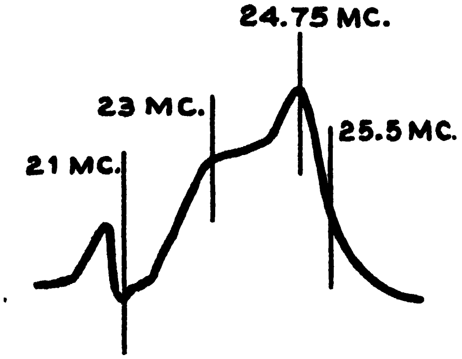


Fig. 16-93.—Over-all curve when adaptor is in 1st video i-f stage.

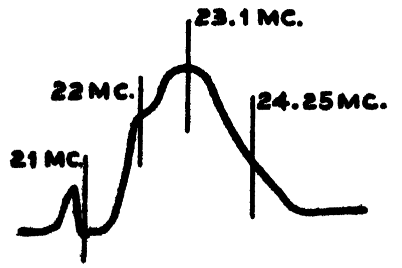


Fig. 16-94.—Over-all curve when adaptor is in 2nd video i-f stage.

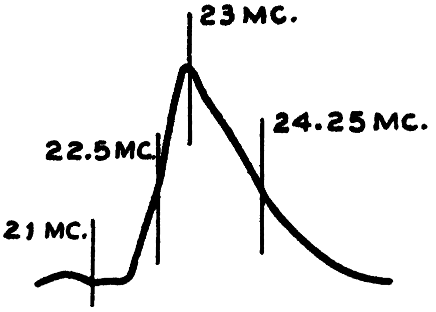


Fig. 16-95.—Over-all curve when adaptor is in 3rd video i-f stage.

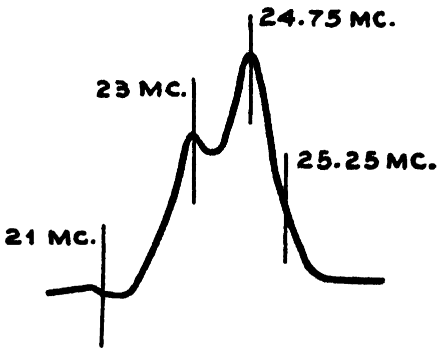


Fig. 16-96.—Over-all curve when adaptor is in 4th video i-f stage.

These measurements were made when a signal was fed into the input of the amplifier and the oscilloscope was connected to the output of the circuit. The effect of the adaptor was checked for the four positions of the device, that is, when it was in the 1st, 2nd, 3rd, and 4th video i-f stage sockets. The effect of increased distributed capacitance upon the over-all response curve is very marked; the frequency of the peak is changed appreciably in two stages. When the adaptor was inserted in the 3rd video i-f stage, the double-peaked curve actually became a single-peaked response.

For the record, the adaptor was constructed in the John F. Rider Laboratories and was made for a miniature tube. Moreover, the receiver on which it was tried employed undercoupled i-f transformers. Naturally, these curves apply to this specific receiver and the effect of such devices will not always be the same, but based upon tests which were made on other receivers, the addition of distributed capacitance in this way during alignment should be avoided. In fact, even lead dress should not be altered after alignment has been completed.

CHAPTER 17

WAVEFORM OBSERVATION IN TELEVISION RECEIVERS

The observation of waveforms of voltages existing in television receivers presents no problems but it does demand the exercise of certain precautions when applying the cathode-ray oscilloscope. Such care will obviate the possibility of misleading conclusions and even possible damage to the display device. Enough background has been given in the sixteen chapters which precede this one so that we can feel free to refer to elements of the oscilloscope and assume that the references are understood.

Test Points in Television Receivers

Anyone who has examined service data concerning television receivers is aware of the numerous waveforms which are offered for reference and which demand the application of a cathode-ray oscilloscope for the diagnosis of troubles. Invariably these waveforms are associated with the vertical and horizontal sweep circuits, the synchronizing system, the video amplifiers, and even the audio-frequency amplifier, although the latter does not receive too much attention. All of these are closely allied to the frequency response of the cathode-ray oscilloscope, especially the vertical amplifier. In view of all that was said about this subject in Chapter 10, it is unnecessary, at this time, to dwell upon the importance of the bandwidth of the .vertical amplifier relative to the appearance of the trace on the screen of the test oscilloscope. However, it is necessary to speak at length about the conditions which surround these tests for they are not always possible under ideal conditions either.

To begin with, the manufacturers of television receivers generally furnish reference information concerning the waveforms which should exist at different points in a television receiver.[1] It is conceivable that an experienced person working on such receivers may elect to observe conditions at places not designated by the receiver manufacturer. If the oscilloscope at hand permits this and the background of the operator is such as to enable proper interpretation of the pattern, all well and good — but if such familiarity is not possessed by the individual, tests other than those recommended by the manufacturer are fruitless and the time might well be saved.

In connection with the reference waveform patterns which are given as guiding information, Fig. 17-1 offers a few of these.

We need not identify them exactly because the subject to be discussed here is of general rather than specific interest. A similarity in pattern conformation at corresponding points exists in all television receivers but the method of producing the trace or the constants of the oscilloscope, which displays the wave later reproduced in a service bulletin, can well result in somewhat different characteristics of the two traces.

Concerning test points, we shall treat these in a general manner. Broadly speaking, signal voltages are to be found at all points in a system which are not at ground potential. In other words, all points in a signal path which are not bypassed to ground will afford a signal voltage wave of some sort. Whether or not this signal can be displayed on an oscilloscope screen depends upon its level relative to the level required by the device which will display the signal. This relates to the sensitivity of the vertical amplifier in the oscilloscope. If it is inadequate with respect to the signal level, display will be impossible. The next controlling factor is the frequency of the signal and the ability of the vertical amplifier in the oscilloscope to pass this frequency. This was discussed in Chapter 10 and will be discussed some more later.

Returning to the test points, it must be realized that any system of waveform examination, regardless of the direction of progress through the system, is the equivalent of signal tracing and the rules of signal tracing apply.[2] In this connection, it might be well to mention that an appreciation of the phase relationships in a vacuum tube and its associated circuit can be very helpful, for quite frequently it enables measurement at a low-impedance point (cathode) with minimum loading by the test equipment, rather than testing at a high impedance such as the control grid, where loading by the test equipment is not only possible, but more than likely.

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[1] Rider Television Manuals Volumes 1 through 8, John F. Rider Publisher, Inc., New York, N. Y.

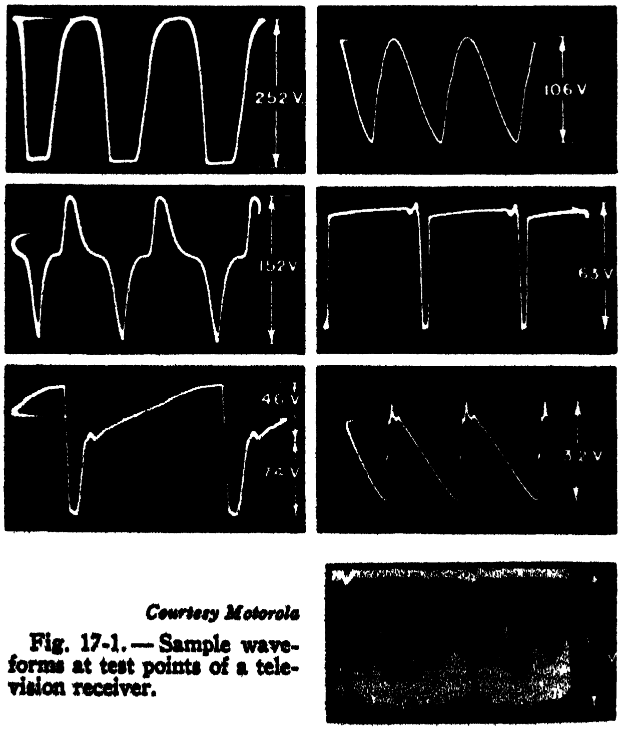
[2] Rider, John F., "Servicing by Signal Tracing," John F. Rider Publisher, Inc., New York, N. Y., 1939.

probable. The same may be said about the plate circuit; in both instances, it is well to understand that signal voltages at the control grid and cathode are in phase — between control grid and plate they differ by 180° and the same is true between the plate and the cathode. This, too, was shown in Chapter 10, but is repeated here because it is directly associated with the observation of waveforms.

Kinds of Waveshapes in Television Receivers

In broadest terms, signal voltages, found in the different parts of television receivers, are of nonsinusoidal character. A few of these are shown in Fig. 17-1. Their exact significance is not of much importance in this discussion. Of far greater concern is the fact that waveform examination in television receivers involves more than just the display of a wave; frequently, it is necessary to determine the peak-to-peak voltage value of the signal. Moreover many of the signal voltages are composites of two or more different voltages which are specially formed so as to perform certain control functions. The amplitude of parts of the wave also is a matter of interest and later we shall deal with the means of measuring voltage amplitudes by means of the oscilloscope.

It is the shape of the wave which is pertinent to performance; the slope of its sides, the rise time, the square corners, the linearity, and the relative amplitudes of the parts. This may be expressed by saying simply its frequency composition. Later in this chapter we shall show the effect of the i-f alignment on the characteristics of control pulses. These details are important because their presence or absence in the trace which appears on the screen is influenced greatly by the characteristics of the oscilloscope.



Courtesy Motorola

Fig. 17-1. — Sample waveforms at test points of a television receiver.

Because of the wide variety of patterns, which one is apt to encounter at corresponding points in different receivers, and since the variations between patterns may be attributed to many factors such as differences in the constants of oscilloscopes, the differences in receiver circuit constants, and the accuracy with which reference illustrations are reproduced, it behooves the individual who has occasion to make such checks to become generally familiar with the kinds of patterns to expect at the different test points in a receiver. It is not necessary that these be memorized; they cannot be, because of circuit arrangements and possible design changes, although, in the end, the deflection systems, whether plates or coils, demand voltage or current waves of appropriate kind.

An understanding of the general principles of operation of these systems cannot help but be very helpful in this connection for it will familiarize the operator with the manner in which currents or voltages are combined to produce special waveshapes. It is not imperative, unless one desires to do so on his own initiative, to remember test points in different receivers. These are fairly well standardized and repeated operations at like points in different receivers will make an indelible impression on one's memory, yet it is to be remembered that service data contains information concerning both the test points and the reference wave at those points. The practice of making changes in a television receiver circuit in the interests of greatest economy in production frequently results in combinations of actions, new tubes, and other changes which can well modify a waveform at a certain point in a circuit. In view of the very many different models and chassis produced by different manufacturers and the attempt to combine functions in tubes, it is well to bear in mind that greatest convenience of operation as well as greatest freedom from possible errors and mistaken conclusions will accrue from the use of the receiver manufacturers' data as contained in service manuals.[1]

As a rule, receiver manufacturers employ wideband oscilloscopes when developing television receiver waveforms for subsequent use in their service manuals. These are capable of showing the voltages in a true light so that the patterns can serve well as reference data. Every so often, however, an oscilloscope with relatively reduced vertical pass-band is employed and the resultant trace of a voltage in the sync, sweep or video system is somewhat inferior in shape. Nevertheless, it will still serve as a guide for, in the final analysis, the importance of the pattern is twofold — 1, the general shape and 2, the equivalent voltage levels. If the observed waveshape during service analysis is a reasonable facsimile of the reference pattern, one part of the requirement has been satisfied. It then becomes a matter of determining if the voltage amplitudes are correct. Of course if the receiver manufacturer's recommendation in oscilloscope equipment is being followed by the service facility, then duplication of the reference waveshape should be possible without any difficulty. The same is true if the service facility's oscilloscope is rated at a higher vertical amplifier bandpass than the original unit employed when the service manual was prepared. But if the service oscilloscope is rated at a much lower vertical frequency response than the instrument used in the manufacturer's service shop, then only an approach to the reference pattern is to be expected. Sometimes this is a close approach and sometimes it is not — depending upon the oscilloscope used — but, generally speaking, it should not be too much of an obstacle in the localization of a fault because the presence of a defect in the system will be productive of a difference in waveshape which is much more radical than would be produced by a mere difference in vertical amplifier frequency response.

In this connection we might recall details pertaining to the performance of uncompensated attenuators in some of the oscilloscopes. These are capable of extreme frequency discrimination as was illustrated in Chapter 10, therefore, care must be exercised to establish that the abnormality which exists in the observed waveshape is a matter of receiver performance and not attenuator action. This is not difficult; all it demands is a variation in the vertical attenuator control position. If it modifies the wave shape in addition to changing its vertical amplitude, it is a reasonable conclusion that much distortion may be attributed to the attenuator control. Under such circumstances it will usually be found that one setting of the control will result in a waveshape which is a fair approach to the desired, provided that the receiver system is performing normally. If it is not so, then no matter what the position of the vertical attenuator, it will be impossible to even remotely approximate the reference waveshape.

Concerning the modification of a wave by the constants of the vertical amplifier, it is, as we described in Chapter 10, a function of the constants of the vertical amplifier in the oscilloscope. As a matter of convenience we show in Fig. 17-2 a series of oscillograms of a 50-microsecond pulse with a pulse-repetition frequency of 4,000 cps as seen on a number of different oscilloscopes, each with a different pass-band in the vertical amplifier. From what has been stated earlier and what appears in these illustrations, it is possible to summarize the effect of the frequency response of the vertical amplifier on the waveshape by saying that the steepness of the sides of the pulse is reduced (sides slant more), square corners are rounded off and flat tops of waves become peaked, as the frequency passband of the amplifier relative to the fundamental frequency of the wave becomes lower.

They are clearly evident in Fig. 17-2 A, B, C, and D. In oscillogram (A) is shown the aforementioned pulse as viewed on an oscilloscope with a 5-Mc vertical-amplifier bandpass. The same pulse displayed on another oscilloscope which has a 500-kc vertical-amplifier bandpass is seen in oscillogram (B). Some deterioration is evident but it does not destroy the utility of the display for maintenance operations. Both shape and amplitude of portions of the wave are as easily determinable from oscillogram (B) as they are from oscillogram (A). For comparison purposes the same pulse is reproduced on the screen of an oscilloscope which has a vertical-amplifier frequency response approximating 100 kc; this is shown in part (C) of Fig. 17-2. The general shape of the pulse has been retained, although slight changes are evident as described earlier in this discussion. Nevertheless it is still possible to ascertain amplitudes of the portions of the wave without too much difficulty. As an extreme case of changes in waveshape due to the frequency characteristics of the vertical amplifier in the oscilloscope, oscillogram (D) is shown. The peak amplitude of the pulse is the same as before, but its square shape identity has been lost completely. The same may be said of the positive parts representing the overshoot.

If the problem were the measurement of the amplitude of the negative portion of the wave, this could be done readily with all four oscilloscopes, but the last oscillogram (D) would be useless as a device which would show that the original pulse was square. Thus the intended purpose of the display plays a very prominent role in determining the adequacy of oscilloscopic equipment.

Additional examples of what may happen to the shape of a pulse as the consequence of incorrect frequency composition are shown in Fig. 17-2E. Here is an illustration of a normal horizontal sync pulse in a television receiver when the over-all response of the i-f system is correct. Three other illustrations of the same pulse under different conditions of frequency versus amplitude in the i-f system also are shown. The loss of the flat top and the sharp corners and a change to a peaked pulse in which it is difficult to define the location of the pedestal — all due to the loss of the high frequencies because of improper over-all frequency response of the receiver — is clearly seen. The other two examples of pulse distortion stem from the presence of excessive high-frequency response with consequent ringing and overshoot on both edges of the wave, and the tilt caused by the loss of low frequencies between 15 and 20 kc, bearing in mind that the fundamental frequency for this pulse is 15,750 cps. The resultant effect on the picture also is indicated.

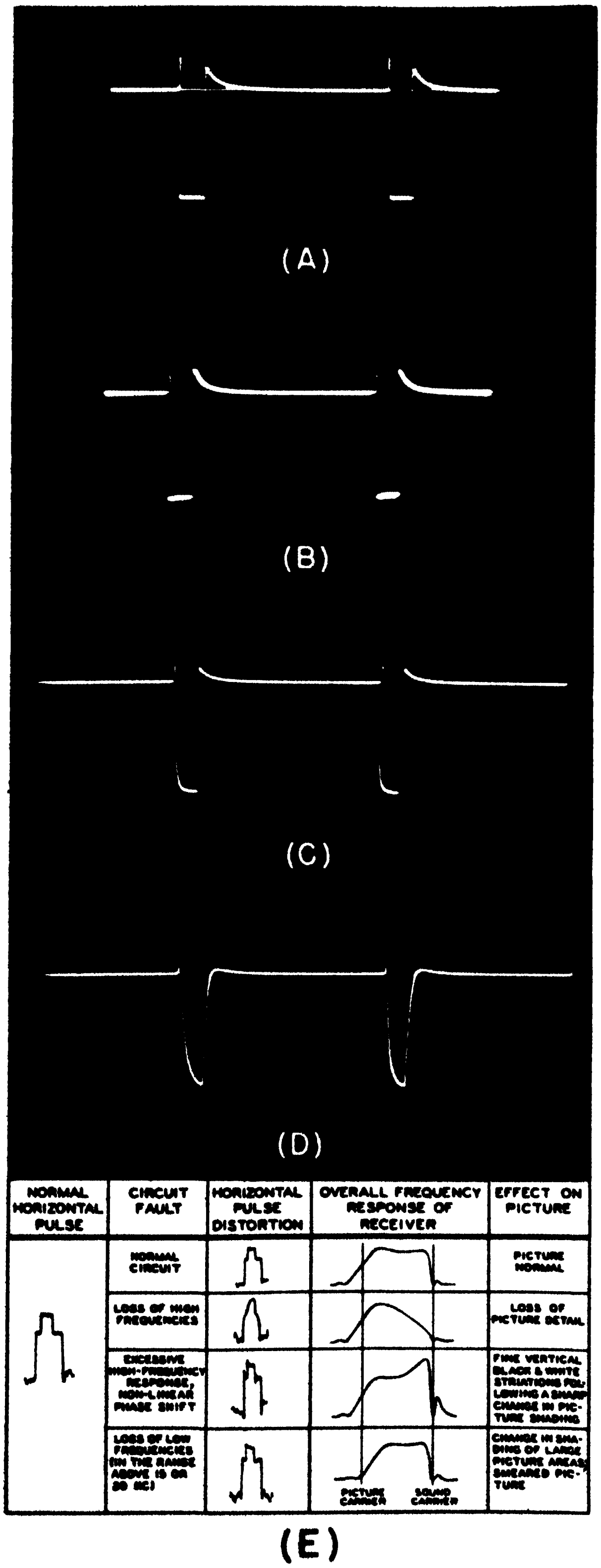
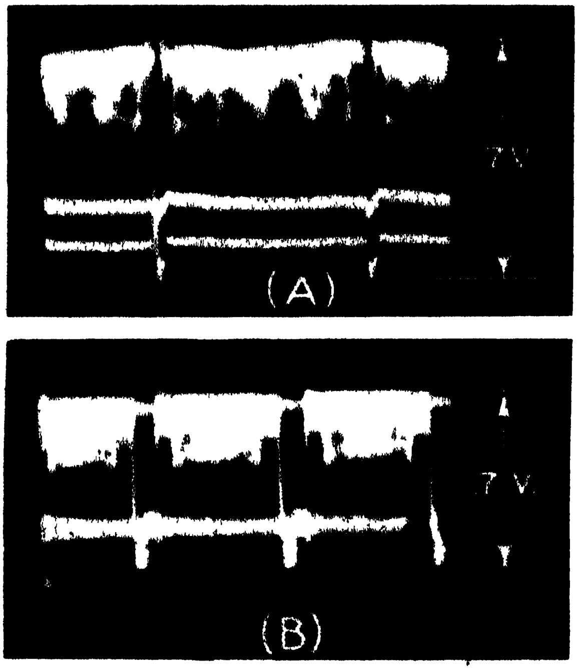


Fig. 17-2.—Responses of different oscilloscopes to a 50-microsecond pulse with a repetition rate of 4,000 per second: (A) 5-Mc vertical-amplifier bandwidth; (B) 500-kc vertical-amplifier bandwidth; (C) 100-kc vertical-amplifier bandwidth; and (D) oscilloscope with input shunted to decrease response to the extent shown. (E) shows the response to a horizontal sync pulse of a television receiver in various stages of alignment.

Limitations Due to Signal Frequency

We have attempted, in earlier chapters of this text, to stress the controlling effects of signal frequency on the application of the oscilloscope, the relationship between the frequency composition of the signal voltage and the pass-band of the vertical amplifier in the test instrument; also the relationship between the frequency range of the time-base oscillator in the oscilloscope and the display of the individual cycles of the signal voltage which is of interest. Recalling the frequency bandwidth of the different oscilloscopes which are available on the commercial market, it is clearly evident that r-f (picture) carriers and the heterodyning oscillator voltage cannot be examined; that is, the individual cycles of these voltages cannot be placed on the screen of all but the very special devices. Even then, and this assumes adequate amplification in the vertical amplifier (which seldom exists), only those picture carriers which are within the 50- to 100-Mc range and those heterodyning oscillator signals which also are within this frequency range may so be viewed. As a matter of fact, it might well be said that because of the added control presented by the frequency range of the time-base oscillator, that r-f and the heterodyning oscillator voltages are not observable on the oscilloscope. The exceptions are rare indeed; only a bare few of the instruments mentioned in this text afford the required vertical-amplifier pass-band and the time-base oscillator frequency range, and even then the level of the received r-f carrier seldom is of sufficient amplitude.

Almost the same thing may be said about the picture and sound i-f signal voltages. The highest vertical-amplifier pass-band found in the great majority of oscilloscopes does not extend beyond several megacycles and this immediately precludes the observation of i-f signals of higher frequency. Even the 4.5-Mc i-f signal is beyond the capabilities of the great majority of devices so that any ideas of examining the individual cycles of these signals might well be forgotten. Moreover the character of the picture and sound i-f signals or the composite i-f in intercarrier receivers is such that it would be impossible to make these patterns stand still long enough for useful study. These comments, therefore, are the equivalent of saying that the r-f, oscillator, and i-f portions of a television receiver are systems which are beyond the display powers of the vast majority of oscilloscopes.



Courtesy Motorola

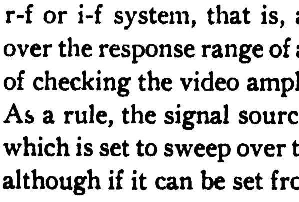
Fig. 17-3.—Video detector outputs: (A) with the sweep frequency on the order of the vertical sync frequency, displaying several complete video frames; (B) with the sweep frequency on the order of the horizontal sync frequency, displaying several complete lines of video.

Of course, rectified versions of these signals are within the display capabilities of the majority of oscilloscopes. Such are the video signals which appear across the video detector output. Two of these are shown in Fig. 17-3A and B, their appearance being different because of the two different sweep frequencies used in the oscilloscope. The former illustrates the pattern for several complete fields whereas the latter is several scanning lines. Because of the nonintegral relationships between the frequencies in the pattern and a single sweep frequency, one oscilloscope sweep setting will not stop all of the elements of the trace, in fact some of the r-f may be lost because of the limited response of the oscilloscope amplifier.

Continuing with the subject of rectified versions of the r-f and i-f signals, the alignment curve or the response curve is another example, although it is representative of what can well be viewed as a singular signal input. As stated in Chapter 16, the alignment or response curve is a pattern of the manner in which the r-f or i-f signal varies in amplitude in a circuit as its frequency is changed. The dominant factor is the rate at which the test signal is varied in frequency. This usually is carried out at either 60 cps or 120 cps, consequently, the amplifier which will pass such a signal need not have a bandpass below say 50 cps and above 130 cps. Of course, we know that this frequency range is within the band of any oscilloscope vertical amplifier, which is the same as saying that any oscilloscope which has a vertical amplifier is adequate for alignment purposes. The required sensitivity is assumed. We shall not dwell upon alignment any longer because of the complete details contained in Chapter 16.

F-M Checking of Video Amplifier

The process used for examining the response of an r-f or i-f system, that is, an f-m signal which sweeps over the response range of a circuit or system is one way of checking the video amplifier in a television receiver. As a rule, the signal source is an f-m sweep generator which is set to sweep over the range of from 0 to 10 Mc, although if it can be set from 0 to 5 Mc, it will be sufficient. This voltage is fed into the video amplifier and the over-all response is determined by locating the rectifying means (which feeds into the vertical amplifier of the oscilloscope) at the output of the video system. Since the rectifying system does not exist at the output of the video amplifier in a television receiver, it must be provided externally and usually is the crystal contained in crystal probes which are part of oscilloscope equipment. Except for the frequency of the sweep generator, the system being checked and the kind of, as well as the location of, the rectifying system, the procedure conforms to conventional response curve tests.



Courtesy RCA

Fig. 17-4.—(A) indicates a double-image video response curve indicating a need for phasing between the horizontal-deflection voltage and modulating signal inside the sweep generator. (B) shows a correct video response curve.

Again the frequency requirement of the vertical amplifier in the oscilloscope is determined by the sweep rate used in the f-m sweep generator. It is either 60 cps or 120 cps, determined by the design of the generator. An example of the "response curve" which may be expected on the screen of the oscilloscope is shown in Fig. 17-4.

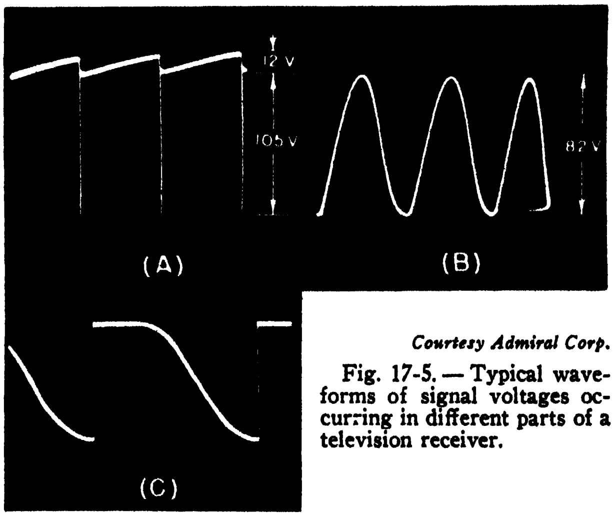
Signal Voltage Levels and Waveshape Dimensions

Waveforms of signal voltages which exist in sweep, sync, and video systems of television receivers are presented in two ways in service data. These are shown in Fig. 17-5A, B, and C. We mean to call attention to the use of reference voltage levels (A) and (B) in some instances and only the wave in others (C). Where the voltage data is not given on the oscillogram, it may be given elsewhere, but if it is absent from the reference data, then the sole purpose of the waveform examination is to find correspondence in terms of waveshape. The presence of the signal voltage of the appropriate waveshape *usually* indicates proper operating conditions; incorrect distribution of voltages in the different tubes of these systems will invariably modify the waveshape to a very marked degree.

It might be well, however, to call attention to several details relating to waveshape. One of these is the advantage of duplicating the frequency conditions indicated in the reference data. By this we mean that the time-base oscillator in the service oscilloscope should be adjusted so as to display the same number of cycles of the signal voltage applied to the vertical-deflection plates as appears on the reference illustration. As a rule, this will result in a pattern which most closely resembles the original, especially if the horizontal dimension gain control or attenuator is arranged to spread the trace so as to cover about three-quarters of the screen face, and in the vertical direction, the trace is spread so that it has an amplitude equal to about one half or perhaps slightly more of the screen height. These are the relative dimensions of most of the reference patterns.

When traces are set up in this manner, the appearance of the pattern is least deceiving when compared with the reference trace. Display of the improper number of cycles or insufficient horizontal amplitude creates an illusion of a difference in the traces which really does not exist. The sides seem steeper and the peaks appear sharper. The true characteristics of the voltage or current wave are not modified by improper display; they just *seem* to look different, and if a comparison is to be made with a reference trace, it is best to set up those conditions which most closely parallel the original. While this is true under all conditions, it is especially significant when applied to traces which do not bear voltage-level labels.

Waveform oscillograms which bear voltage-level values as in Fig. 17-5A and B warrant comment. To begin with, the cathode-ray oscilloscope is suitable for the measurement of signal voltage levels with results in absolute terms, but it is important to understand that this can be accomplished under specific conditions only. The measurement of the trace height and conversion into signal voltage at the input demands certain specific procedures if the determination is to be meaningful. Before any measurement can be made, the screen of the oscilloscope must be calibrated. Sometimes receiver manufacturers, who show voltage levels on their waveform oscillograms, specify a certain oscilloscope and state specific control positions. These control positions are the equivalent of calibration under certain circumstances, but it might well be stated definitely that the calibration of every oscilloscope is a necessity.



Courtesy Admiral Corp.

Fig. 17-5. — Typical waveforms of signal voltages occurring in different parts of a television receiver.

There is still another way of treating such voltage values when they appear on oscillograms. The amplitudes of the parts of the wave can be considered in relative terms or in absolute terms. The previous examples treated them in the latter manner. Much information can be developed by simply examining the relative heights of the parts of the wave without determining the absolute voltage levels of each. For example, in the wave shown in Fig. 17-5A, the sawtooth portion has an amplitude slightly more than one-tenth of the negative pulse, and conversely, the negative pulse has an amplitude which is just under nine times that of the sawtooth portion. These values may be determined by the simplest means of linear measurement; a ruler is a convenient method.

Such relative amplitude measurements are useful for while they will not indicate whether or not the signal values are what they should be, the general conditions found in television receivers are such that, if the relative amplitudes of the parts of the wave are correct, the chances are that the operating conditions, in the circuit where the measurement is made, are correct. Such information can lead to proper conclusions of circuit conditions.

Occasionally, the over-all amplitude of the wave suffers because of circuit parameters although the relative amplitudes of different parts of the wave are not modified. This may tend to limit the degree of control offered by a circuit, but in most instances, a defective state in a stage or in a system will modify the shape of the wave as well as its amplitude. Occasions when only the amplitude is changed are rare indeed.

Voltage Calibration of Oscilloscopes

A convenient method of using the oscilloscope to measure a-c signals is the substitution method. A signal whose amplitude can be varied is substituted for the video or sync signal to be measured. This signal is usually a 60-cps signal. The amplitude of the 60-cps signal is then adjusted until its pattern on the oscilloscope is the same amplitude as that of the signal to be measured. The peak-to-peak voltage of the 60-cps signal is then measured, either by a meter or a calibrated control. The amplitude of the video or sync signal will then be the same as that of the 60-cps signal.

A circuit such as shown in Fig. 17-6 can be used to supply the 60-cps signal which is substituted for the signal being measured. The transformer used will depend on the range of voltages to be measured. For instance, a heater transformer with a 6.3-volt secondary will supply a peak-to-peak signal of 6.3 × 2.83 or almost 18 volts peak to peak at the highest setting of the potentiometer, which can be about 1,000 ohms in this case. If higher voltages are used, the value of the potentiometer should, of course, be higher, in order not to exceed the voltage rating of the potentiometer or the current rating of the transformer.

The 60-cps signal can either be measured by an a-c voltmeter, as shown in Fig. 17-6, or the potentiometer can be calibrated in terms of peak-to-peak volts. If a meter is used, it should be remembered that most a-c meters read rms volts and it is necessary to multiply the readings by 2.83 to obtain peak-to-peak values. If a potentiometer is used, it must be calibrated by hand unless it has a very good linear characteristic, in which case a linear dial can be used.

When the 60-cps signal is substituted for a signal being measured, the gain control and the attenuator setting on the oscilloscope should not be changed. If desired, the sweep frequency of the oscilloscope can be reset in order to obtain 2 or 3 cycles of the 60-cps signal, but this is not necessary as the only factor that affects the measurement is the amplitude of the signal, and this will be the same regardless of the oscilloscope sweep frequency. In this respect, it might he helpful if the horizontal gain control is turned to zero so that no horizontal deflection is obtained. The signal being measured and the 60-cps signal will then both show up as straight vertical lines whose over-all heights can be readily compared.

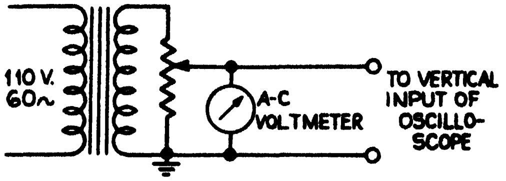


Fig 17-6.—Circuit for voltage calibration of an oscilloscope.

The commercial market affords devices for voltage calibration purposes. These are described elsewhere in this text as oscilloscope accessories. Whether one such is used or the simple arrangement of Fig. 17-6 is employed, the process of calibration is virtually the same. As stated in Chapter 10, some oscilloscopes employ self-contained voltage calibrators. These are very convenient, of course, but the substitution method is still used, although being part of the oscilloscope, the operation is accomplished with very little time lost even when each calibration involves a substantially different voltage value.

Typical Voltage Calibration

Suppose, for example, that we assume a trace 4 inches high on a 5-inch screen, or a trace 2 inches high on a 3-inch screen and say that both of these represent a calibrating input voltage (at input of vertical amplifier) of 50 volts rms. This corresponds to a peak-to-peak voltage of 50 × 2.83 or 141 volts, which we may call 140 volts in round numbers. Measurement in peak-to-peak values is made for a definite reason. First of all, the oscilloscope trace indicates peak-to-peak values; second, the conversion from rms to peak-to-peak values for sine waves is simple because the relationship is constant. That is not so for complex waves. The ratio between the rms and peak-to-peak values differs for each kind of waveform. It is rather difficult to determine this ratio, and if this is not known, it is not possible to compare the amplitudes of waves of different form by merely comparing their rms values. The use of peak-to-peak values eliminates this difficulty. It has also been found that peak-to-peak values are more significant in analyzing the function of various circuits in television receivers.

The voltage values mentioned above were arbitrarily selected and can just as readily represent 5 volts peak to peak or 500 volts peak to peak. It all depends upon what is the desired range of voltage measurement, as determined by the work to be done, and also by what is the setting of the vertical-amplifier attenuator. Relative to the latter, it is very important to understand that, once the calibration has been completed, it holds as long as the vertical-amplifier attenuator is kept *constant* in its position. If it is varied, the calibration no longer will hold. This means that voltage measurements with the oscilloscope may require a series of calibrations so as to suit the range of voltages to be measured. Logically, a calibration for measuring a signal voltage of several hundred volts peak to peak with a trace 2 or 3 inches high will be very unsatisfactory for measurements of just a few volts. It will be impossible to read the voltage level from a trace of such a small dimension as would result from an attenuator setting which accommodates voltages several hundred times greater.

Referring once more to the sample voltage level previously mentioned, let us assume that the trace is 4 inches high and that it represents an rms voltage of 50 volts and a peak-to-peak voltage of 140 volts. Then 140 volts equals 4 inches and each inch of deflection equals 35 volts peak to peak. If a linear crosshatch scale is placed across the face of the tube, the kind which may show 5 squares per inch, then each square vertically represents 7 volts peak to peak. We repeat that this holds as long as the vertical attenuator is maintained constant in its position for that particular calibration. If, instead of a 4-inch trace on a 5-inch tube, the same voltage is displayed by a 2-inch trace on a 3-inch tube and a similarly marked scale is in front of the screen, then each square would correspond to a peak-to-peak voltage of 14 volts.

Having completed the calibration, the waveform to be measured is applied by connecting the input of the oscilloscope to the point where the wave of voltage exists and the proper trace is displayed by adjusting the time-base oscillator frequency. *The vertical-amplifier attenuator is not touched!* Assuming that the proper value of voltage was considered in the original calibration, the height of the wave to be examined does not exceed the original 4 inches. If it does, it will introduce no difficulties in voltage measurement provided that it does not extend beyond the curvature of the screen near its ends. Let us further assume that the trace on the screen is the wave (C) in Fig. 17-5A which we reproduce as Fig. 17-7.

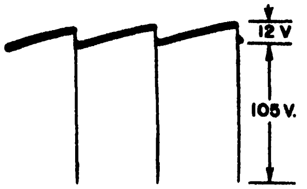


Fig. 17-7.—In this sweep waveform the voltage of the sawtooth portion is seen to be much smaller than that of the negative portion.

This wave might be described as consisting of two parts, the sawtooth portion which has an amplitude of its own and the negative pulses which extend below the point where the sawtooth begins. As far as the cathode-ray beam is concerned, it treats the wave as a single entity, but because of the calibration of the trace in the vertical direction or due to the calibrated screen on the face of the tube, it is possible to determine the peak-to-peak voltage value of any portion of the wave.

Considering the total voltage between the top and bottom limits of the wave, it is seen to be the sum of 12 + 105 or 117 volts and this can be determined by noting the length of the trace vertically between the squares of the calibrated screen, or by measuring the over-all height of the trace and converting it to voltage on the basis of 35 volts peak to peak per inch. If the linear scale is in front of the tube screen, then the voltage amplitudes of the two parts of the wave can be determined readily by counting the squares which include the respective parts of the wave. The sawtooth portion will include slightly less than 2 squares whereas the negative pulse, beginning at the bottom of the sawtooth wave, will be located within 15 squares. This is shown in Fig. 17-8.

If the linear scale is not used, then it becomes necessary to measure the respective portions of the trace and, with an over-all length of 4 inches corresponding to 140 volts peak-to-peak, the sawtooth portion will have a vertical length of slightly more than 0.25 inch and the negative pulse will be about 3 inches long.

The calibration of the oscilloscope for voltage measurements is in no way limited by the frequency re-

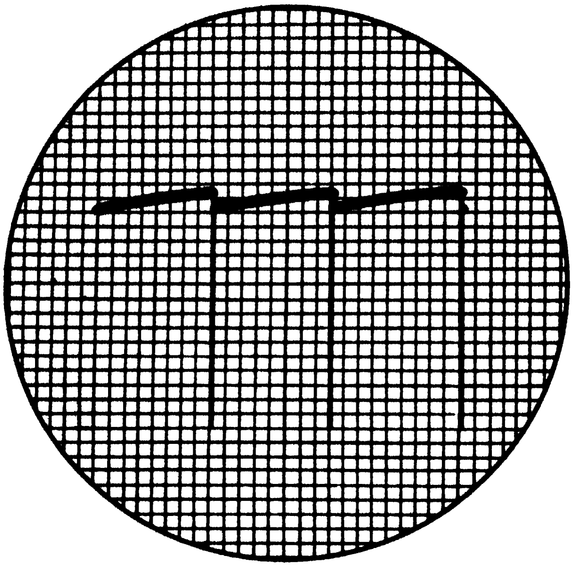


Fig. 17-8.—Fig. 17-7 seen against a calibration grill.

sponse characteristics of the device. The latter is strictly a function of the design of the amplifier system and the same may be said of the frequency range of the time-base oscillator in the instrument. Regardless of what these ranges may be, the vertical deflection (also the horizontal deflection when necessary) may be calibrated in terms of signal voltage at the input to the system. The procedure is the same for all oscilloscopes although those devices which are equipped with self-contained calibrators usually are accompanied by special instructions which apply to the specific equipment. Changing the settings of the vertical gain control or attenuator will change the calibration since the control determines the magnitude of signal fed into the amplifier, therefore, the dimensions of the trace per volt input. This is not to be confused with the deflection factor or sensitivity rating of the tube. Neither should it be confused with the sensitivity rating of the amplifier-deflection system combination as expressed in the oscilloscope manufacturer's literature. This figure of merit states the maximum sensitivity with the attenuaator set to zero attenuation or 1:1 ratio or the gain control wide open.

Range of Voltages Measurable with the Oscilloscope

The signal voltages which can be measured in absolute levels (after calibration) with the oscilloscope bear definite limits on the low end as well as on the high end. The minimum values are determined by the oscilloscope sensitivity rating as given in the manufacturers' literature. Usually this is stated in terms of a 1-inch deflection for a certain number of volts or fraction thereof rms. Voltages less than this value may be measured if a trace dimension of less than 1 inch is considered usable. Sometimes it is and sometimes it is not, depending on the nature of the application.

At the high end, the limiting factor is the voltage breakdown rating of the equipment which comprises the input system of the oscilloscope. Each oscilloscope bears such a rating and it should be observed. Usually it is given in rms values and sometimes it is given in d-c values. Whichever it may be, it is important to remember that the d-c rating corresponds to the peak a-c rating and *not* the rms rating. Thus if an input system is rated at 600 volts d.c. and it is to be applied across an a-c circuit, the maximum safe rms value is 600 × 0.707 or roughly 420 volts rms. The reasoning behind this is that the d-c voltage rating corresponds to the peak a-c rating and the multiplying factor of 1.414 must be applied to the rms value in order to arrive at the peak value.

In view of the voltages which are encountered in television receiver systems, oscilloscopes designed for such uses usually contain input systems which will accept about 1,000 volts d.c. (707 volts rms) across the input circuit. Again we might remark that the safe input voltage varies in different instruments, therefore, it behooves the operator to determine the operating capabilities of his instrument and to use it accordingly. It is unsafe to take for granted that the constants of one device are duplicated in the products of another manufacturer, or even in another model oscilloscope produced by the same manufacturer.

Effect of Screen Diameter on Voltage Calibration

It is unsafe to consider the deflection sensitivity rating of the cathode-ray tube as the basis of voltage calibration. On the other hand, the full sensitivity rating of the amplifier-deflection system combination is very limited in use because a reasonable amount of signal voltage at the input will drive the trace beyond the limits of the tube screen and the trace dimension then cannot be used for determining signal levels. It, therefore, becomes necessary to decide the range of voltage measurement which will be encompassed by the calibration. (This requirement is greatly simplified by self-contained calibrators.)

For example, if the oscilloscope manufacturer's rating of maximum sensitivity is used, the highest voltage measurable with the gain control wide open is determined by the screen diameter, assuming about 90 per cent of the screen surface to be the useful diameter. This means that a typical rating of 0.05 volt rms per inch for a 5-inch tube will allow measurement of voltages up to about 0.25 volts. On the other hand, if the voltage to be measured approximates 100 volts rms, the calibration will have to be made with such setting of the vertical-amplifier gain control that the system sensitivity is say 50 volts rms per inch. A trace dimension 0.5 inch high then will correspond to 25 volts rms and a trace dimension about 4.5 inches high will establish the other limit of 225 volts rms.

Effect of the Oscilloscope Input Circuit on Waveform

The constants of the input system of the vertical amplifier in the oscilloscope are capable of materially influencing the shape of the signal being examined. This is true especially in those circuits wherein a resonant condition is responsible for the shape of the signal. The reason for the action is the capacitance of the oscilloscope input system which is added across the circuit in the receiver. Sometimes the effect is negligible in view of the high amount of capacitance already present at the point where the oscilloscope is connected; the added capacitance being relatively small. Or the input to the oscilloscope is via the attenuator probe, wherein the capacitance is held to from 5 to perhaps 10 $\mu\mu f$. But when the oscilloscope input system adds from 20 to 30 $\mu\mu f$ across a shunt- or series-peaked circuit as in video amplifiers, the change in waveshape becomes a matter of chance, being dependent upon the frequency of operation. The higher the frequency at which the peaking takes place, the greater is the effect of the capacitance loading. Fortunately, this problem is limited, in the main, to the video amplifier where it always is advisable to locate the oscilloscope input at such a point in a stage where the impedance is lowest.

From the practical viewpoint, the problem is not as serious as a theoretical discussion would indicate because the reference patterns shown in service data take cognizance of the loading. These patterns are actual practical conditions with the oscilloscope in place and all that is needed is to match or approach the type of pattern shown. Moreover, the conditions of test usually are stated and, while it is possible that the oscilloscope used for the observation may not be the exact one employed by the manufacturer when the reference data were prepared, the difference in input capacitance between different brands of probes or different brands of direct input circuits is seldom sufficient to introduce problems.

The input resistance or impedance of the oscilloscope, direct or through a probe, is capable of displaying a loading effect which will tend to reduce the amplitude of the signal voltage. Once more, however, reference data already take into account such loading so that unless the reference material was prepared much higher than that used for the maintenance work, the effect of loading will not be serious, at least not as far as the ability to locate a defective condition is concerned.

Relative to the use of low capacitance probes as the input channel to the oscilloscope, it must be realized that these attenuate the signal, usually by a factor of 10, so that when checking waveforms, the introduction of the probe will very materially reduce the over-all amplitude of the signal fed into the oscilloscope. This degree of attenuation may be sufficient to make the display useless for study purposes, in which case, use of the regular input will be required and the effect on the waveshape recognized and allowed for in the analysis. Two effects may be noted; these are the possible loss of some high frequencies due to the increased capacitance and possible loss in over-all amplitude due to the circuit loading caused by the comparatively low input resistance presented by the oscilloscope input circuit. The latter effect is not usual, however, unless the measurement is made across a very high impedance circuit.

Phase of Voltage Waveform

As explained in Chapter 10, the phase of the voltage trace which appears on the screen of the oscilloscope is a function of the number of stages in the vertical amplifier of the testing device, or to say it more broadly, it is a function of the specific design of the oscilloscope amplifier. Interest in this condition lies in the possibility of confusion when comparing a reference trace with a test trace when the two are of opposite phase. Sometimes the oscilloscope contains a means of inverting the trace on the screen through 180°. Many oscilloscopes do not contain this facility, in which case a slight problem arises. It will be necessary to correlate the two traces by visualizing the trace on the test oscilloscope screen turned through 180°. The equivalent of this is to examine the reference trace from the rear of the sheet on which it is printed, if this is possible. Simply turning the trace upside-down is not the answer because this is not the same as turning the image through 180°. Examples of this are shown in Fig. 17-9A, B, and C. Illustration (A) is assumed to be a reference illustra-

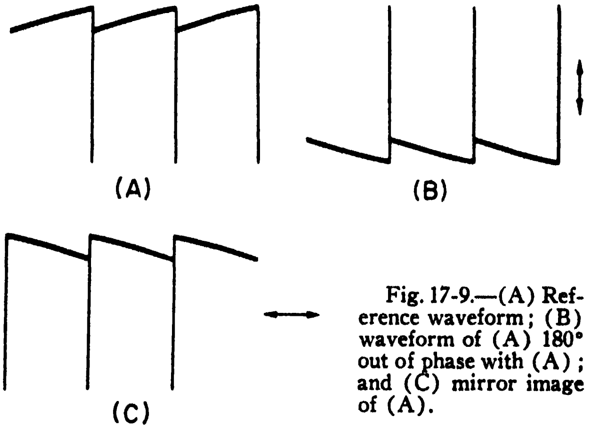


Fig. 17-9.—(A) Reference waveform; (B) waveform of (A) 180° out of phase with (A); and (C) mirror image of (A).

tion whereas illustration (B) is the wave as seen on the test oscilloscope screen. A comparison of (A) and (B) shows that the two differ by 180° in polarity. Illustration (C), on the other hand, indicates that waveform (A) has been turned over in the horizontal direction as shown by the double-headed arrow. This is not the same as turning it over in a vertical plane as shown by the double-headed arrow in waveform (B). The latter is the equivalent of the 180° phase change which is accomplished in the amplifiers. A change in orientation of the pattern so that it corresponds to mirror symmetry, shown by oscillograms (A) and (B) in Fig. 17-9, is the correct one.

Waveforms in Low-Voltage Power Supplies

In view of the discussion contained in Chapter 19, it is unnecessary to dwell upon this phase of waveform observation in television receivers in this chapter. These procedures, which are described in general terms, are applicable in their entirety to the low-voltage power supplies in television receivers.

Required Frequency Range of the Oscilloscope

Fundamental frequencies associated with operations in the sync, sweep, and video systems of a television receiver usually are 60 cps and 15,750 cps. But due to the shapes of these waves, a wide range of frequencies are involved. The background for this condition was discussed in detail in Chapter 10. In the meantime we might say that certain minimum frequency requirements exist in the oscilloscope which is used for waveform observation in such equipment. At the low end of the range, a low limit of say 20 to 30 cps is adequate and this figure is attainable in virtually every oscilloscope. In fact there is no problem relative to the lowest frequency. It is at the high end of the pass-band of the vertical amplifier that problems arise. For general service operations, we can say that an oscilloscope with a top frequency of 300 kc for the vertical amplifier is entirely suitable. Of course it is better if the upper limit of frequency is 1.0 Mc or even higher for reasons which are given in Chapter 10, but the lower limit stated above is entirely satisfactory. At the low end of the vertical-amplifier bandpass, the lower the limit the better, but a low limit of 20 to 30 cps is entirely satisfactory. As a rule the commercial equipments offer a lower range than this.

Square-Wave Testing

The most conclusive information about the merit of a video amplifier is obtained by feeding a square-wave voltage to the amplifier input and observing the waveform of the voltage at the amplifier output. The square-wave test is made as follows:

1. Set up the equipment as shown in Fig. 17-10.

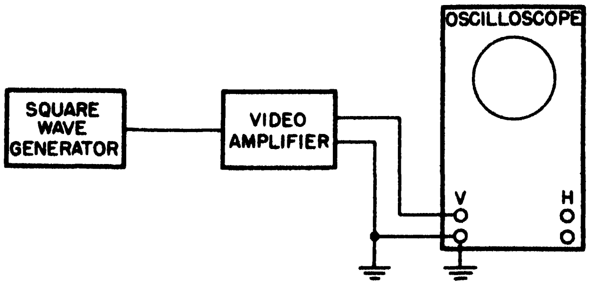
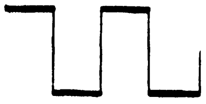
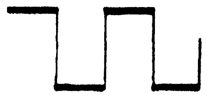
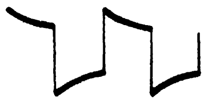
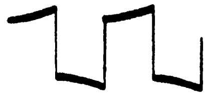
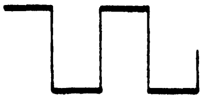
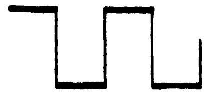
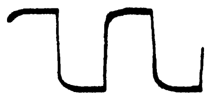
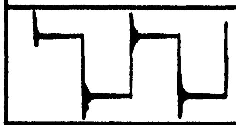
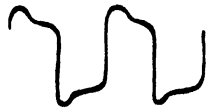


Fig. 17-10.—Block diagram of circuit for square-wave testing.

2. Adjust the square-wave generator so that a 60-cps wave is fed to the video amplifier (keep the generator output low so that the video amplifier will not limit and give an erroneous pattern). The wave observed on the oscilloscope will give much information about the low-frequency response of the video amplifier. A chart showing some of the curves which may be seen, together with the video-amplifier defects responsible, is shown in Fig. 17-11.

3. Set the square-wave generator so that a 25-kc wave is obtained (as before, the generator output should be kept low to avoid overloading of the video amplifier). Whenever possible, the attenuating probe should be used with the oscilloscope for this test, since its capacitance shunting the video amplifier output is

INPUT TO VIDEO AMPLIFIER	SHAPE OF OUTPUT WAVE SEEN ON C-R SCREEN	EFFECT ON TELEVISION PICTURE	DEFECT	CHECK THESE CIRCUIT COMPONENTS
SQUARE WAVE (ABOUT 60 CYCLES) 		PICTURE NORMAL	NO DEFECTS GOOD LOW-FREQUENCY RESPONSE AND NEGLIGIBLE PHASE SHIFT.	
		GRADUAL CHANGE IN PICTURE SHADING FROM TOP TO BOTTOM OF PICTURE. (THIS EFFECT CAN BE MINIMIZED BY D-C RESTORER ACTION)	LEADING LOW-FREQUENCY PHASE SHIFT. (USUALLY ACCOMPANIED BY A LOSS OF LOW-FREQUENCY GAIN)	COUPLING CAPACITORS, SCREEN AND CATHODE BYPASS CAPACITORS, LOW-FREQUENCY COMPENSATION CIRCUITS, SCREEN AND GRID RESISTORS
			LAGGING LOW-FREQUENCY PHASE SHIFT USUALLY CAUSED BY OVER-COMPENSATION.	
SQUARE WAVE (ABOUT 25 KC) 		PICTURE NORMAL	NO DEFECTS. GOOD HIGH-FREQUENCY AND TRANSIENT RESPONSE.	
		PICTURE DETAIL IS POOR; SHARP CHANGES IN PICTURE SHADING ARE "FUZZY."	POOR HIGH-FREQUENCY RESPONSE.	PEAKING COILS, LOAD RESISTORS, LEAD DRESS OF PEAKING COILS AND COUPLING NETWORKS.
		FINE VERTICAL BLACK & WHITE STRIATIONS FOLLOWING A SHARP CHANGE IN PICTURE SHADING.	EXCESSIVE HIGH FREQUENCY RESPONSE & NON-LINEAR TIME DELAY, ALSO HIGH FREQUENCY CUTOFF MAY BE TOO SHARP.	PEAKING COILS, LOAD RESISTORS, DAMPING RESISTORS SHUNTING PEAKING COILS, LEAD DRESS.
		WHITE BORDER FOLLOWING A BLACK-TO-WHITE TRANSITION; BLACK BORDER FOLLOWING A WHITE-TO-BLACK TRANSITION.	EXCESSIVE OR NOT ENOUGH MID-FREQUENCY RESPONSE & NON-LINEAR TIME DELAY	PEAKING COILS, LOAD RESISTORS, DAMPING RESISTORS SHUNTING PEAKING COILS, LEAD DRESS.

Courtesy RCA

Fig. 17-11.—Tabulated information on the square-wave testing the frequency response of video amplifiers.

less than when the direct probe is used. When the video amplifier has a low-impedance output, such as would be obtained when the last stage is a cathode follower working into a low-impedance line, this precaution need not be observed.

Information about the high-frequency response of the video amplifier may be obtained from the curves observed on the cathode-ray tube. Fig. 17-11 shows some of the effects, together with the characteristics which produce them.

CHAPTER 18

TRANSMITTER TESTS

The application of the cathode-ray oscilloscope to the testing and adjustment of transmitters involves measurements of amplitude, frequency, and phase characteristics. All of these have been discussed previously in this book. The manner in which each of these is important is determined by the nature of the measurement — by the specific section of the transmitter which is under test. No new principles are involved, so that much of what has been said in preceding chapters will be found applicable to the pertinent parts of transmitters. It is, of course, important that the basic operation of the transmitter be understood for the oscillograms which are developed by means of the cathode-ray oscilloscope are nothing more than pictorial evidence of the conditions of operation existing in the system being examined. They indicate the presence of correct operating parameters or the existence of incorrect relationships of voltage, current, resistance, or impedance.

The labels associated with the oscillograms describe the electrical conditions which may give rise to the waveshapes shown. Quite frequently several related factors are responsible; one of several may be responsible but it is impossible to select one of these to the total exclusion of all others because one operating parameter may influence the status of another. Since only a limited number of oscillograms can be shown conveniently, several of each type are offered as guides so as to satisfy the greatest number of variations which an operator may experience. No pretense is made that more could not have been given, nor that other kinds of transmitters, or places in transmitters could not have been investigated. Unfortunately, space limitations did not allow the inclusion of more oscillograms. Subsequent revisions of this text will offer elaborations of what is contained in the first edition. The same is true of the kinds of transmitters which are covered. It so happens, however, that the transmitter arrangements selected and most completely represented are those which find more prominent use, therefore, the tests illustrated by oscillograms will serve the greatest number of readers. At least it is hoped that such will be the case.

Basic Circuit

Fig. 18-1A and B shows the two basic systems for making oscilloscope measurements on amplitude-modulated transmitters. The process involves feeding the r-f signal (modulated or unmodulated as the case may be, invariably the latter) to the vertical-deflection system. This may be directly to the vertical-deflection plates without benefit of the vertical amplifier, or to the vertical-deflection plates via the vertical amplifier. The controlling influence is the frequency of the r-f signal relative to the frequency pass-band of the vertical amplifier. Generally speaking, the most commonly used arrangement is to feed the r-f signal directly to the plates and this applies to all frequency ranges.

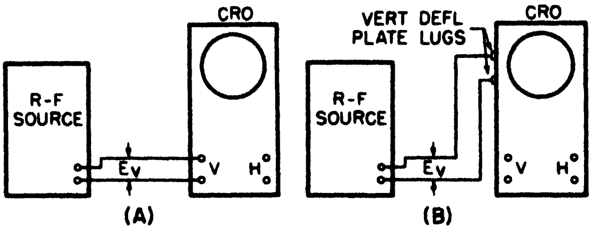


Fig. 18-1.—(A) If the frequency of the r-f signal is within the frequency range of the vertical amplifier of the oscilloscope, voltage *Ev* may be fed to the input of that amplifier. The method of securing such signal voltages conveniently is shown in Fig. 18-7. (B) If the frequency of the r-f signal is beyond the response limits of the vertical amplifier of the oscilloscope, the signal should be fed directly to the vertical-deflection plates by either of the two methods shown in Fig. 18-7. A much stronger signal will be required when feeding directly to the deflection plates than when feeding the signal to the vertical-amplifier input. This calls for more turns on the coupling coil or closer coupling.

The above reference does not mean that only the r-f signal voltage may be applied to the vertical-deflection system. This is one kind of measurement, that involving the output of the transmitter, but it must be realized that examination of any other signal in the transmitter proper may be investigated in the various ways which have been described in preceding chapters. The cathode-ray oscilloscope does not change its identity simply because it is applied to a transmitter. The different sections of the instrument serve the numerous purposes which were described in Chapter 10.

It so happens that the most prominent tests made on a-m transmitters involved the matter of modulation, or the modulated carrier. This is so because the conditions of operation in the modulator stage and in the final r-f output stage determine, to the greatest extent, the behavior of the transmitter; whether or not it radiates a signal which is free from spurious emanations and affords utmost intelligibility. Improper conditions in other parts of the test transmitters also are shown, but in the main, they revolve around these two parts of the system. As stated previously, the most frequently used type of a-m transmitter was selected for the portrayal of conditions. This is the high-level plate-modulated system, using a Class B modulator and a Class C r-f amplifier.

To obtain a number of significant oscillograms directly, three transmitters were set up and operated under different conditions. In all cases except where otherwise noted, the modulating signal was a wave of about 400 cps fed into the speech amplifier.

Most of the oscillograms shown are of three types. One type is the carrier with or without modulation, wherein the linear time base in the oscilloscope is responsible for the horizontal deflection. Only the r-f output from the transmitter is fed to the oscilloscope. This kind of trace is frequently identified as a block pattern. Because the sweep frequency is very low in comparison to the modulating frequency (if it is used) the pattern appears as a rectangular block of light because neither the individual cycles of the carrier or the envelope of carrier amplitude variations corresponding to the modulation can be seen. The presence of modulation is visible in such patterns in the form of light of greater brilliance, which traverses the block. Examples of such patterns appear in Figs. 18-2 through 18-4.

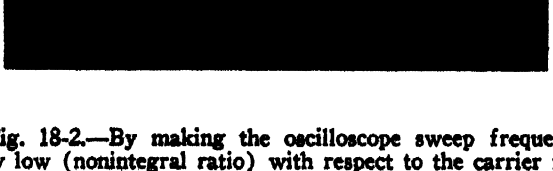


Fig. 18-2.—By making the oscilloscope sweep frequency very low (nonintegral ratio) with respect to the carrier frequency which is without modulation, the carrier signal voltage trace will appear as a rectangle of light on the screen. If the trace is free of streaks (horizontally) the signal is relatively free from harmonics. If it is badly streaked, the harmonic content is high. Do not confuse these irregularly located streaks with the modulation bands as shown in Figs. 18-3 and 18-4.

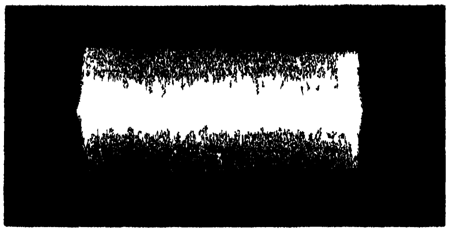


Fig. 18-3.—If the oscilloscope sweep is adjusted to a frequency which is low relative to the modulating frequency (nonintegral ratio) the modulated carrier will appear as a rectangular trace of light. The single streak of bright light which divides the pattern in half indicates 100 per cent sine-wave modulation.

The increased brightness of the trace responsible for the appearance of the lines in the modulated patterns is due to the positioning of the negative or downward modulation peaks. The higher the percentage of modulation, the closer are these points of maximum trace intensity (slowest beam velocity across the screen) to each other. With 100 per cent modulation, they form a single narrow line. For lower percentages of modulation, these lines are located farther apart. The absence of such a line in the trace indicates the absence of modulation, although it must also be mentioned that such a block pattern of a carrier high in harmonic content will produce a streaked pattern wherein the lines of greater intensity will not uniformly divide the pattern in the horizontal direction.

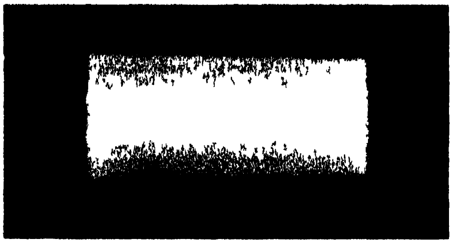


Fig. 18-4.—With conditions the same as in Fig. 18-3, the two bright lines traversing the trace and dividing the pattern into three equal areas indicate 50 per cent sine-wave modulation. Complex wave modulation could well produce several such horizontal bands of light and still correspond to the same amount of modulation.

The second type of oscillogram is the one which illustrates the modulated-wave envelope or modulated carrier. This is a plot of the changes in carrier amplitude or r-f voltage with time. In this case, too, the r-f signal from the transmitter output is fed to the vertical-deflection system and the linear time base or sweep in the oscilloscope is adjusted to some submultiple of the modulation frequency. After proper synchronization, the pattern will remain stationary. The trace is typified by Fig. 18-5A and Fig. 18-6. The third main category of oscillograms is the trapezoid; in this case, the

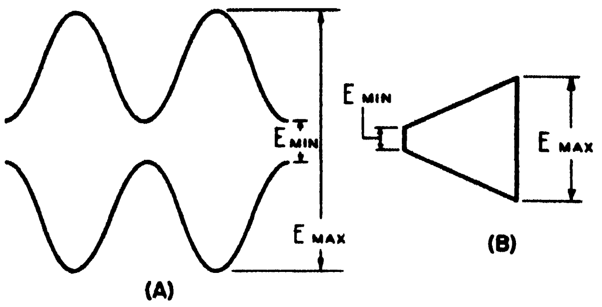


Fig. 18-5.—(A) Modulated-wave pattern, (B) trapezoidal type trace.

changes in r-f (carrier) voltage are plotted against changes in modulating voltage. The r-f voltage is applied to the vertical-deflection system of the oscilloscope and the modulating voltage is applied to the horizontal-deflection system (usually to the horizontal amplifier input) through appropriate circuits. The conformation of the trapezoid appears in Fig. 18-5B.

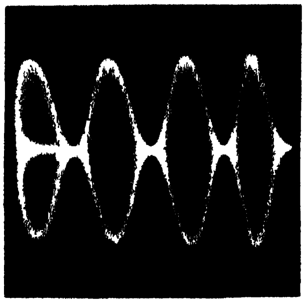


Fig. 18-6.—Oscillogram of the wave envelope showing 100 per cent sine-wave modulation of the carrier. This trace will be referred to in subsequent illustrations.

The trapezoidal pattern can reveal many details of transmitter operation which are not clearly evident in the wave envelope type of trace, therefore, it is to be preferred for numerous tests. It is simpler to develop, requires no critical oscilloscope time-base or synchronizing adjustments, and is equally suitable for sine- or complex-wave modulation. The wave envelope type of trace can be developed for complex-wave modulation but unless all the modulating components are in harmonic relation with the sweep frequency, the trace will wander and will not be suitable for study purposes.

These three kinds of traces may be developed with speech modulation, and they are very interesting to observe. The block and wave envelope types of patterns will be seen to rise and fall in height with modulation, whereas the trapezoid will be noted to expand laterally from a single vertical line to a trapezoid or to a triangle which is lying on its side, dependent on the character of the modulation process.

Examination of the individual cycles of the carrier are not readily possible, although if the frequency is low enough relative to the time-base oscillator frequency range, a maximum about 50:1, the manner in which the different amplitudes of the carrier cycles form the wave envelope when the carrier is modulated will be discernible. Such display is not particularly valuable in connection with transmitter testing; the usual wave envelope and trapezoid presentations are entirely adequate.

Precautionary Measures

It is well, right at the outset, to issue warnings concerning the possibility of danger when making waveform tests in transmitters which involve high operating voltages in the r-f and modulator systems, for that matter, throughout the transmitter. Circuits used to feed r-f and modulator output signal voltages to the oscilloscope are shown later in this chapter. The utmost precautions must be used in the handling of connecting leads which make direct connection between the oscilloscope and portions of the transmitter. It is always best if the r-f voltages are inductively coupled rather than capacitively coupled. If it must be the latter, then the operating voltage ratings of the coupling capacitors should be at least twice the highest voltage possible in the transmitter at the point where the signal is being tapped off.

Only well-insulated connecting leads should be used and every connection must be solid and well made. Shielded wire is preferable so that the shield may be grounded and thus avoid complications due to feedback. Connections to high-voltage circuits must be insulated against flashover to ground. Connections should not be made with the high voltage applied to the transmitter. If connections are made to the deflection plates via pin-jacks or terminal boards, adequate separation must be provided between the point of contact and near-by rounded points so as to protect against flashover because it may do irreparable damage to the oscilloscope. Care must be exercised to observe the voltage limits specified in the oscilloscope equipment specifications. Whenever possible, r-f voltages, applied to the deflection plates or to the vertical amplifier input, should be obtained via inductively coupled circuits as shown later in Fig. 18-7.

When feeding signals from the modulator stage to the horizontal amplifier as shown in Fig. 18-7, care must be exercised in the selection of capacitive and resistive elements against voltage breakdown or flashover. It should be remembered that signal voltages, applied to the input circuit of oscilloscope amplifiers, need not exceed 1 volt, as a rule, in order to afford a satisfactorily dimensioned trace.

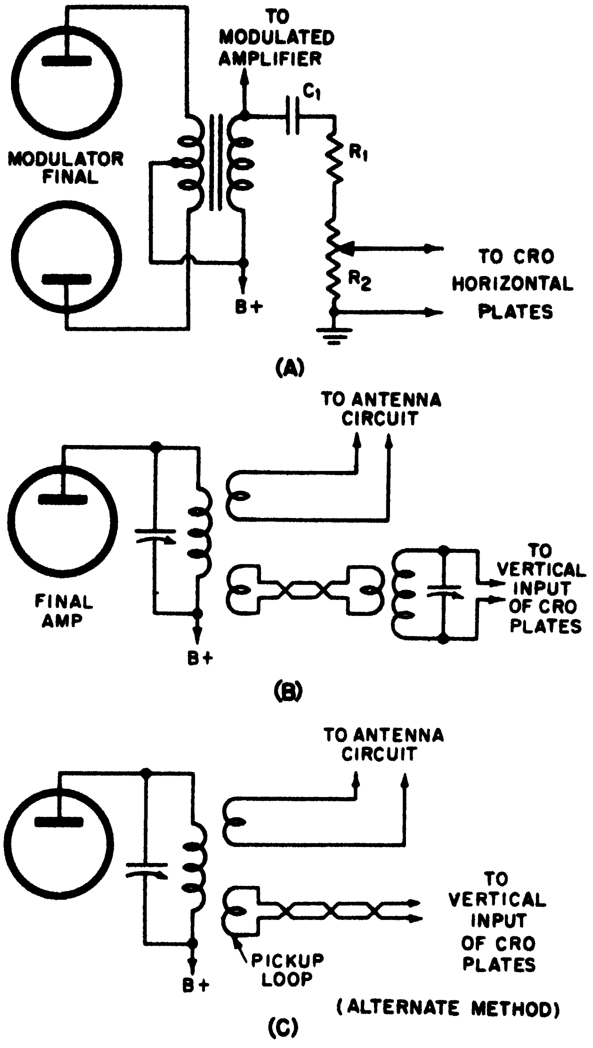


Fig. 18-7.—(A) Capacitance-coupled voltage divider connected across the modulator output transformer secondary feeds signal voltage of controllable amplitude to horizontal amplifier of oscilloscope for trapezoidal pattern. Impedance of the shunt path should be at least 10 to 20 times the impedance presented by the r-f output tube. The voltage required for the oscilloscope amplifier need not exceed several volts. This determines the relative values of the fixed and potentiometer sections of the divider. Voltage breakdown precautions previously discussed should be observed. (B) Link coupling feeds energy to the circuit tuned to the operating carrier frequency (usually the same as the plate tank) and the signal is then fed to the vertical-deflection system. The r-f signal level and its character are determined by the degree of coupling between the link and plate tank and by the condition of resonance in the tuned coupling circuit. The rotor of the tuning capacitor should be grounded and so should the shield of the twin coax line which connects the link to the tuned circuit. Coupling between the link and plate tank should be as loose as possible consistent with adequate pickup. Use of a tuned coupling link is not preferred because it can result in wrong information. (C) This method is preferred to the system shown in Fig. 18-7B since the trace developed on the screen of the oscilloscope is a more accurate representation of the signal fed to the antenna. If the transmitter employs a tuned antenna coupler, the link may be coupled to it, or to the plate tank, whichever affords the most convenient and best arrangement. The link need not contain more than a few turns and the connection can be twin coax with the shield grounded. Sometimes the link may have to be replaced by a tuned circuit resonant to the operating frequency.

Amplitude Modulation

There are two main types of patterns that may be obtained in amplitude-modulation tests: Modulated-wave envelope and trapezoidal patterns.

A modulated-wave pattern, or envelope, is shown in Fig. 18-5A. The internal sweep of the cathode-ray oscilloscope is adjusted to a submultiple of the modulating frequency. For example, using 400-cps modulation, the sweep frequency would be 400/4 = 100 cps to show 4 cycles on the tube screen, as in Fig. 18-6 The vertical deflection is obtained using the circuit of Fig. 18-1.

The height of the peak above the horizontal axis plus the depth of the peak below it will give the peak-to-peak dimension $E_{\max}$, in Fig. 18-5, and the trough-to-trough dimension, across the horizontal axis, is $E_{\min}$. The figure compares carrier patterns with the trapezoidal (to be discussed shortly). The percentage of modulation is, then

$$\text{Percentage} = \frac{E_{\max} - E_{\min}}{E_{\max} + E_{\min}} \times 100$$

where $E_{\max}$ and $E_{\min}$ are measured in any convenient units. A typical oscillogram of a carrier modulated by a sine wave is shown in Fig. 18-6.

When a trapezoidal pattern is wanted, the audio voltage used for modulation is also used for horizontal deflection. Since only a portion of this voltage is required, some sort of voltage reduction is used. A practical method is indicated in Fig. 18-7A. *R1* and *R2* form a high-resistance voltage divider. The signal level output is controlled by adjusting the movable contact on *R2*. The blocking capacitor *C1* should be 0.1 μf or

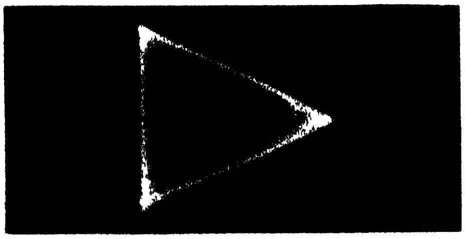


Fig. 18-8.—Example of a trapezoidal trace corresponding to 100 per cent sine-wave modulation wave envelope shown in Fig. 18-6. Note symmetry of pattern, the straight sides and the sharp apex. This indicates linear modulation and is an example of proper operation of the modulator and final r-f stage. It indicates that positive and negative peaks of modulation are alike; that the output current is proportional to the plate voltage on the Class C r-f output stage. Examples of dissimilar conditions appear later.

larger and have a d-c voltage rating twice the maximum potential in the circuit. The total resistance, of *R1* + *R2*, should be about 2,000 ohms per volt of modulator output (0.02 megohm per volt). The r-f signal may be picked up by the arrangement in Fig. 18-7B or 18-7C.

The percentage-of-modulation formula previously given also applies to the trapezoidal case. A typical trapezoidal pattern is given in Fig. 18-8.

Developing the Trapezoidal Pattern

The vertical displacement of the spot on the tube screen is a function of the modulated r-f signal; the horizontal displacement corresponds to the audio voltage from the modulator or some other selected point. The trapezoidal figure establishes the relationship between the two signals visually. Since the rate of amplitude modulation of the carrier is equivalent to the modulating frequency and the identical audio signal is used for horizontal deflection, the pattern and the sweep do not need synchronizing.

Fig. 18-9 shows the development of the trapezoidal pattern. It indicates, at (A), two cycles of the modulated wave envelope and one cycle of the audio voltage, with the resultant trapezoid structure. The numbered points on the envelope of the modulated wave correspond to those on the audio wave, and on the developed trapezoid. The negative peak of the modulated wave corresponds to the negative peak of the audio wave. Since both halves of the modulated wave are identical, corresponding positions of the envelope may be similarly numbered.

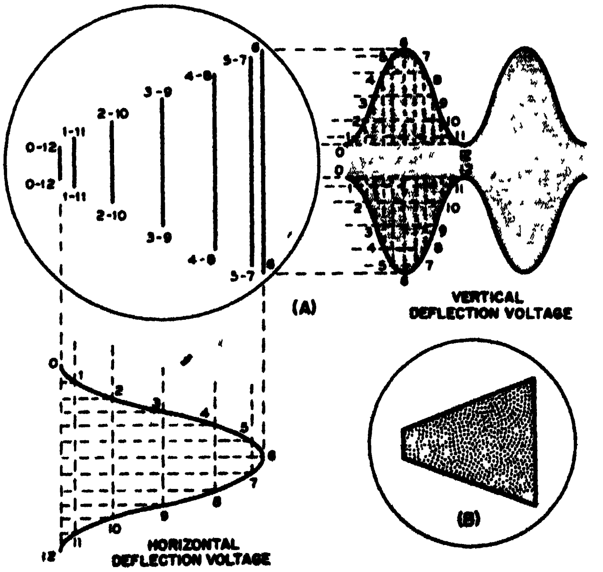


Fig. 18-9.—(A) Development of the trapezoidal pattern; (B) the final trapezoidal pattern.

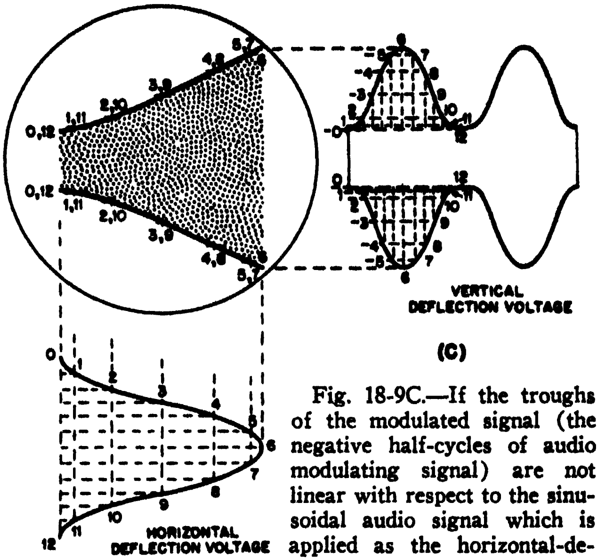


Fig. 18-9C.—If the troughs of the modulated signal (the negative half-cycles of audio modulating signal) are not linear with respect to the sinusoidal audio signal which is applied as the horizontal-deflection voltage, the trapezoid will curve inward.

Displacement of the spot across the screen in one horizontal direction is accomplished by the change in voltage between points *1* and *6* of the audio wave. The return trace of the spot in the horizontal direction is accomplished by the change in voltage between points *6* and *12* of the audio wave, so that the completed pattern really consists of two traces, but since the horizontal displacement of the spot is occurring at a fast rate, the pattern appears as a single image. When the spot is in position *0* along the *X*-axis, it is being displaced along the vertical or *Y*-axis by the voltage difference between *0-0* of the modulated wave at that instant; a single line, therefore, is evident upon the screen. It is to be remembered that the vertical motion of the spot takes place at the carrier frequency, and the peak of each wave of the carrier is determined by the audio modulation. As the spot moves horizontally toward the other end of the screen, that is, toward position *6*, the peak amplitude of the modulated signal voltage is also increasing at exactly the same rate, so that as the spot reaches positions *1, 2, 3, 4*, etc., along the horizontal axis, the modulated carrier voltage also reaches points *1, 2, 3, 4*, etc., and the displacement of the spot in the vertical direction is determined by the peak voltages between points *1-1, 2-2, 3-3, 4-4*, etc., of the modulated wave.

It is evident that as the spot moves in the horizontal direction, from *1* to *6*, vertical displacement will also take place for the various peak values of the modulated carrier. At any instant, lines, such as are shown for spot positions *1* and *2,* appear upon the screen.

As the spot moves to *6,* representing the limit of the sweep in one direction, the vertical displacement corresponds to the modulated peak voltage between points *6-6* upon the modulated curve. As the spot starts moving back across the screen and reaches point *8* upon the horizontal sweep-voltage curve, the vertical displacement is that due to the peak voltage between points *8-8* upon the modulated wave. This point corresponds with points *4-4*. As the spot reaches point *12* along the horizontal sweep-voltage curve, which is the audio voltage curve, the vertical displacement is that due to the peak voltage between *12-12.* This corresponds to points *0-0,* and the cycle repeats. The result is a stationary pattern of the type shown in Fig. 18-9B. It is to be noted that the form of the modulating wave has no effect upon the pattern. The only requirement for the pattern to be a straight-sided trapezoid is that the amplitude of the carrier be linear with respect to the modulating voltage. By this is meant that the change of carrier amplitude, with respect to an unmodulated carrier, is proportional at any instant to the amplitude of the modulating viltage. If the amplitude of the carrier is not linear, the trapezoid will not have straight sides. This is indicated in parts (C) and (D) of Fig. 18-9.

It may happen that the modulation envelope may differ in phase from the audio signal, especially if the audio connection is to a point other than the modulator output. This means, of course, that there is a time difference between the signal and the envelope; for this reason, a return trace will not fall exactly upon the forward trace. Fig. 18-10 is an illustration. Here the envelope lags the modulation slightly; it reaches its peak at the instant *6,* while the modulating voltage is already decreasing, having passed its peak at the instant *5.* Fig. 18-11 illustrates the development of a trapezoid pattern showing where the r-f is completely cut off (overmodulated) for part of the cycle. The method outlined for Figs. 18-9 and 18-10 is applied to Fig. 18-11.

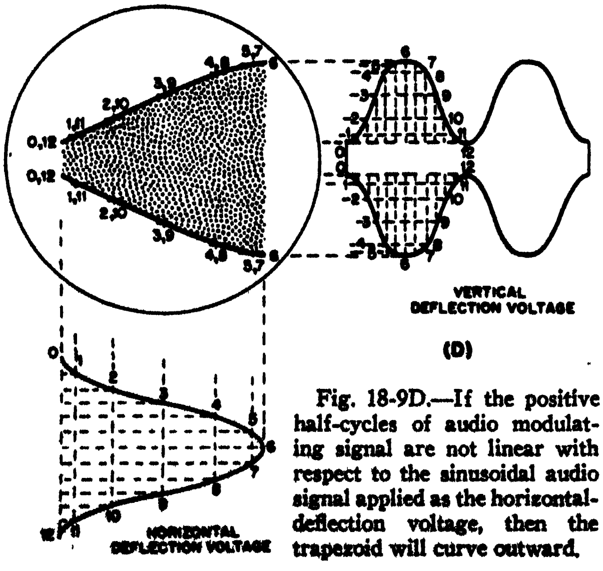


Fig. 18-9D.—If the positive half-cycles of audio modulating signal are not linear with respect to the sinusoidal audio signal applied as the horizontal-deflection voltage, then the trapezoid will curve outward.

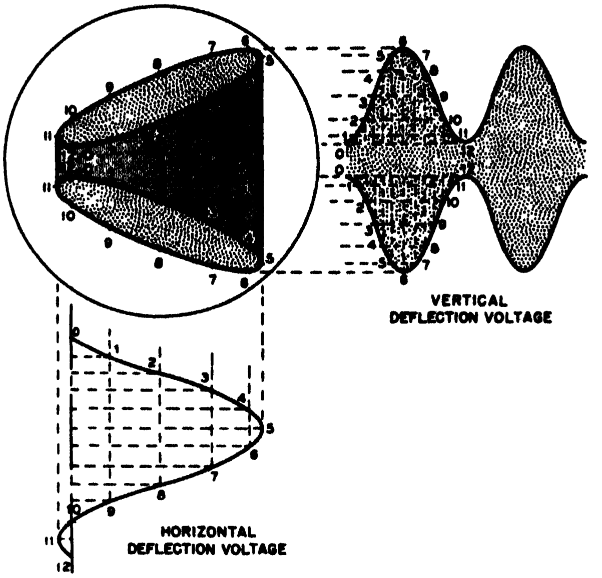


Fig. 18-10.—Development of a trapezoidal pattern when the phase of the modulator output and r-f voltage differ from that which existed in Fig. 18-9. The elliptical sides of the trace should be borne in mind because they may be encountered in practice as shown later. They do not reflect improper operation of the transmitter (Fig. 18-9), but are shown as the background for subsequent illustrations wherein the trace on the oscilloscope involves a phase relationship between r.f. and audio at the oscilloscope which differs from the conditions actually existing in the transmitter.

As a matter of fact, the upper and lower boundaries of the trapezoid (or approximate trapezoid) are a pair of Lissajous figures[1] formed by the modulation envelope and the modulating voltages, with the space between them occupied by the r-f trace. While a nonsinusoidal signal input will not affect the linearity of the sides of the trapezoid, it will introduce a series of bright vertical bands in the pattern, corresponding to the irregularities present in the audio signal. This is the result of nonuniform horizontal motion; as the spot moves slower and faster, the individual r-f waves are closer or wider, and, accordingly, the pattern appears brighter or dimmer.

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[1] See Chapter 12 for a discussion of Lissajous Figures.

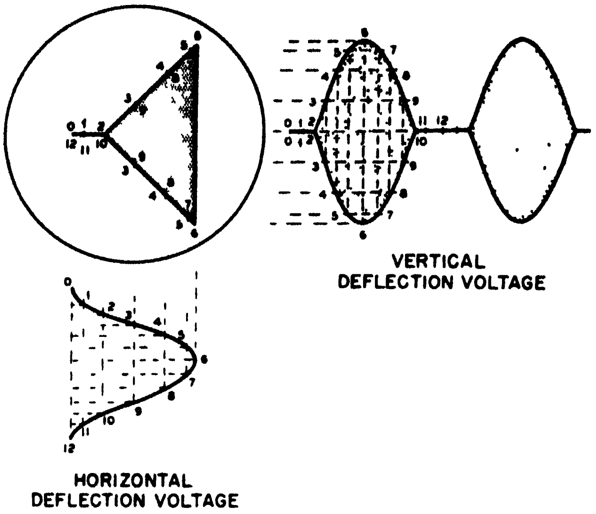


Fig. 18-11.—Development of a trapezoidal pattern when overmodulation exists. Note that the carrier is cut off completely during a part of the negative cycle of the audio voltage. This is shown by the horizontal line without any vertical deflection. Different degrees of overmodulation determine the extent of this line relative to the remainder of the trace. Note the straight sides of the trapezoid which indicate that modulation is linear until the instant when the audio exceeds the level required for carrier cutoff.

Symmetry of modulation can be determined by examination of the wave envelope or trapezoidal patterns. In wave-envelope patterns, the positive and negative peaks will be alike in form and amplitude, if there is symmetry. The carrier height (no modulation) must be midway between the positive and negative audio heights. When judging trapezoidal patterns for symmetry, a horizontal reference line drawn from the midpoint of the E_{min} amplitude should bisect the E_{max} amplitude. Both halves of the pattern, above and below this horizontal reference line, should be alike. Modulation percentage is found in the same manner as used for the carrier pattern.

A-M Transmitter Output Oscillograms

In view of the details concerning oscilloscope operation which were given in previous chapters, we shall devote little space here to that subject. A few very pertinent matters will be commented on, but for the general conditions, we refer the reader to preceding chapters especially Chapters 9 through 17. A general precaution is to make certain that r-f energy intended for the vertical-deflection system is not coupled into the horizontal system and vice versa; if this happens, the patterns will contain details which can well lead to confusion and erroneous conclusions.

It would be well to reiterate here that, unless otherwise stated, sine-wave modulation was employed. Since the conditions of test may not always be the same and the patterns seen on a test oscilloscope may be of various kinds although the tests may be of the same type, a number of oscillograms are shown for certain kinds of tests. These will tend to create the greatest familiarity with the subject at hand and most likely satisfy most of the test conditions which will be used for the checking of amplitude-modulated transmitters.

Typical Proper Conditions of Operation

Relative to the display of transmitter oscillograms, Fig. 18-12 illustrates a modulated-wave envelope trace when the sweep in the oscilloscope is not properly adjusted; that is, the frequency of the sweep is not an integral submultiple of the modulating frequency.

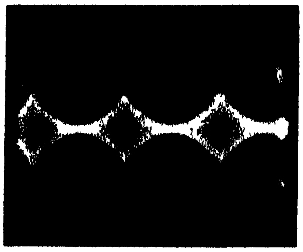


Fig. 18-12.—Example of a wave-envelope trace when the oscilloscope sweep frequency is not an integral submultiple of modulation frequency. The pattern will be in motion. (See references to sweep adjustment in Chapter 10.)

Proper adjustment of the oscilloscope time-base frequency control and the synchronizing control results in the stationary pattern illustrated in Fig. 18-13. For more details on such adjustments, refer to Chapter 10.

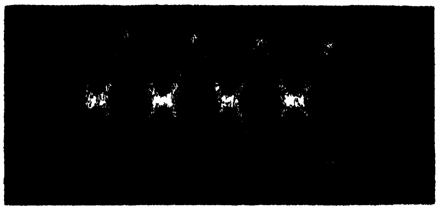


Fig. 18-13.—Example of a modulated-wave envelope trace when the oscilloscope sweep frequency is an integral submultiple of the carrier modulating frequency. When properly synchronized, the trace will be stationary and can be studied. The vertical dimension of the trace is a function of the coupling between the r-f source and the link system which feeds the modulated signal to the oscilloscope deflection system.

For examples of transmitter operation when conditions in the modulator and final r-f stages are correct see Fig. 18-14. Obviously these traces are predicated upon the existence of proper conditions ahead of the final r-f stage and ahead of the modulator stages. Variations of these appear later and are separately classified.

Fig. 18-14.—(A) Example of a modulated-wave envelope showing the carrier modulated 30 per cent by a sine wave. (B) Example of a trapezoidal trace for conditions described in Fig. 18-14A.

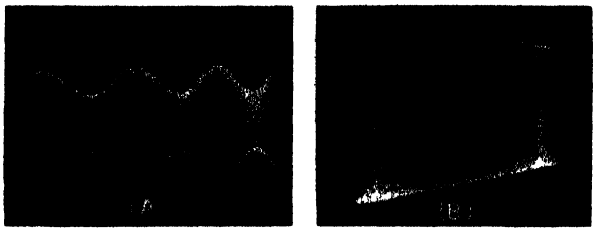


Fig. 18-14.—(C) Example of a modulated-wave envelope when the carrier is modulated 50 per cent by a sine wave. (D) Block pattern of an r-f carrier modulated 50 per cent by a sine wave (Fig. 18-14C). The oscilloscope sweep frequency was very low in comparison with the modulating frequency for this trace.

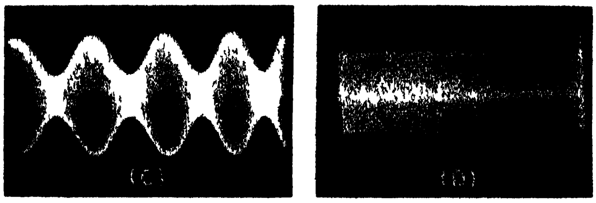


Fig. 18-14.—(E) Example of a modulated-wave envelope showing approximately 90 per cent modulation by a sine wave. (F) Trapezoidal pattern corresponding to conditions described in Fig. 18-14E.

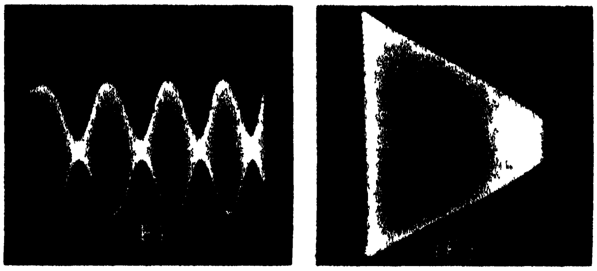
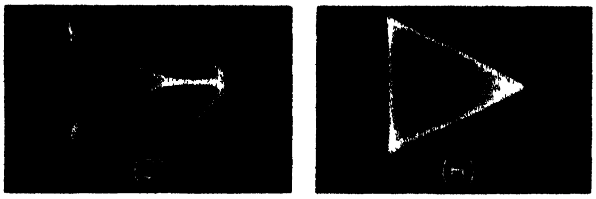


Fig. 18-14G and H.—Trapezoidal traces corresponding to the 100 per cent modulation shown in Fig. 18-6. The phase between a-f and r-f components is different. Note that the difference between (G) and (H) is a function of the test arrangement and is not evidence of a change in the character of the modulated wave. Without modifying the transmitter, pattern (G) can be changed to that of (H) by simply changing the point which feeds the audio modulating signal to the horizontal amplifier of the oscilloscope; that for (G) was taken from the output of the modulator whereas for (H), it was secured from the input to the speech amplifier.



(R-F PLATE VOLTAGE)

Fig. 18-15.—(A) Four cycles of the modulated-wave envelope to be used as the standard of comparison. This trace corresponds to proper plate voltage applied to the r-f final stage with respect to the modulator output stage. (B) Here is the same trace changed in appearance and amplitude by a reduction in plate voltage applied to the r-f final stage. Flattening of both positive and negative peaks is visible, although the effect on the upward modulation is more pronounced. The plate voltage applied is still sufficient to prevent overmodulation.

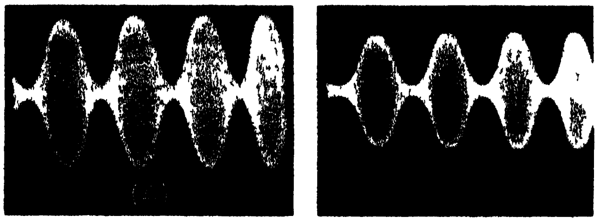
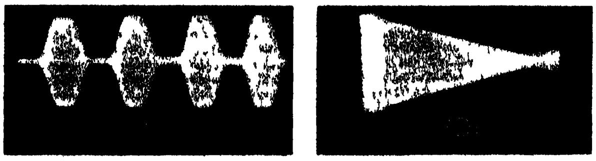


Fig. 18-15.—(C) The trace in Fig. 18-15B is still further modified by a greater reduction in plate voltage. The positive peaks have been flattened to a much greater extent and overmodulation is very pronounced. A good deal of linearity prevails in the modulation between the limits set by the plate voltage. (D) The trapezoidal trace corresponding to Fig. 18-15C shows the flattening on the upward swings in modulation, the overmodulation at the downward swings and the linearity between. Obviously a reduction in modulator output voltage or an increase in r-f plate voltage will remedy the situation.



(INSUFFICIENT EXCITATION—EXCESSIVE BIAS —OVERMODULATION)

Fig. 18-15.—(E) Oscillogram of the modulated wave when the r-f excitation applied to the r-f final stage is insufficient. This prevents a rise in output r-f current proportional to the increase in plate voltage. Under some conditions, the amount of linearity of modulation in the downward directon may exceed that shown here. (F) The trapezoidal pattern corresponding to Fig. 18-15E. As is evident, the amplifier modulates nonlinearly for almost the complete modulating cycle.

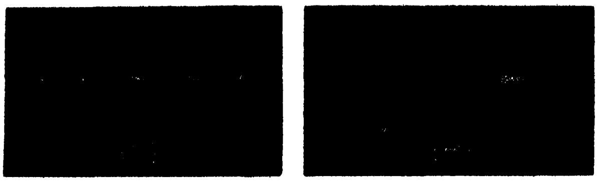


Fig. 18-15.—(G) The nonlinearity in modulation due to improper r-f final bias (excessive) is clearly evident, although 100 per cent modulation of the carrier is being accomplished. (H) This is the trapezoid corresponding to the modulated-wave envelope shown in Fig. 18-15G. The sides of the trapezoid are concave throughout their length indicating the absence of linearity in modulation throughout the cycle.

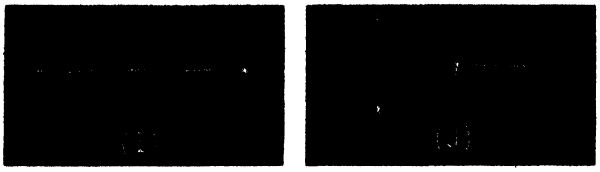


Fig. 18-15.—(I) Oscillogram of the modulated-wave envelope when the excitation is insufficient and the r-f final bias is excessive, but the modulator output is normal. Virtually the entire negative audio cycle is absent in the modulated wave. Note the nonlinearity during modulation in the upward direction. (J) Here is the trapezoidal pattern corresponding to the wave envelope shown in Fig. 18-15I.

Fig. 18-15.—(K) Oscillogram of a modulated wave when r-f excitation to the final is very low and the modulator output is such as to cause some overmodulation. The inability of the output current to rise with plate voltage applied to the r-f tube is evidenced by the very flat-topped wave. (L), (M) Here are two examples of the trapezoidal pattern corresponding to Fig. 18-15K. Pattern (L) is for the correct phase conditions at the oscilloscope, whereas trace (M) indicates a difference in the r-f voltage—audio voltage phase relationship in the transmitter and at the oscilloscope.

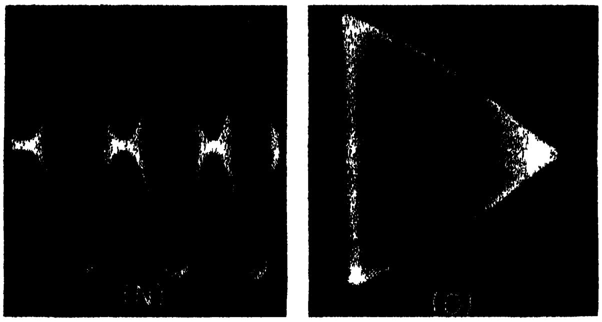


Fig. 18-15.—(N) An oscillogram of a carrier modulated 100 per cent by a sine wave when the modulation is not linear on the upward swing. Compare with Fig. 18-15G. (O) The trapezoidal trace corresponding to Fig. 18-15N. The curvature of the sides of the trapezoid is the exact opposite of that shown in Fig. 18-15H. It is interesting to note that the curvature of the sides of the trapezoid is related to the linearity of the upward and downward modulation. Sides which curve inward (Fig. 18-15H) indicate nonlinearity during downward modulation; sides which curve outward indicate nonlinearity during upward modulation. Nonlinearity during both directions of modulation can result in a trapezoid with sides which curve outward over a portion of the trace and inward over another portion of the trace.

(MISMATCH BETWEEN MODULATOR AND R-F FINAL STAGE)

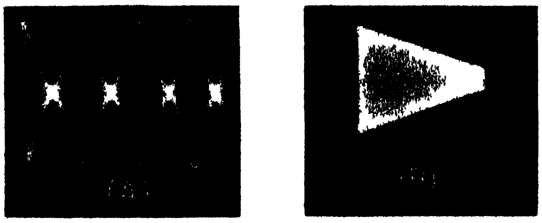


Fig. 18-16.—(A) Oscillogram of modulated-wave envelope for proper match between modulator and r-f final stages. Compare with oscillograms which follow. (B) Oscillogram of trapezoidal pattern corresponding to Fig. 18-16A.

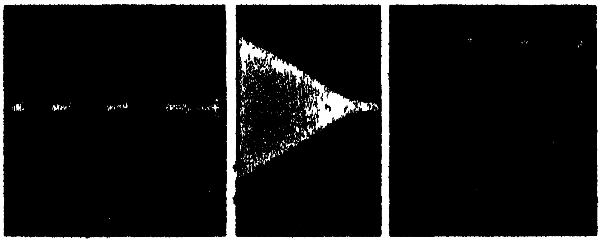


Fig. 18-16.—(C) Oscillogram of modulated-wave envelope corresponding to Fig. 18-16A when the modulator stage feeds into an impedance which is higher than required. Without any change in input signal to the modulator stage, the voltage output from the modulator obviously is higher than when conditions are proper and overmodulation of the carrier results. (D) Trapezoidal trace corresponding to Fig. 18-16C. (E) Oscillogram of audio voltage out of modulator as the consequence of the change in load impedance applied to the modulator stage. Compare with Fig. 18-16H.

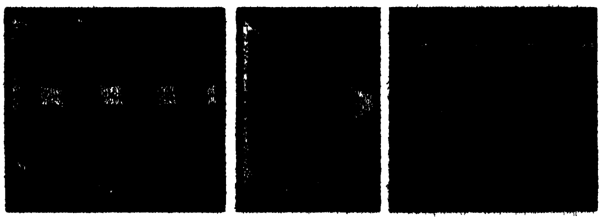


Fig. 18-16.—(F) Oscillogram of modulated-wave envelope when the load on the modulator was less than required. (Compare with Figs. 18-16A and C.) The effect of the incorrect matching in this direction was two fold. It reduced the modulating voltage output and reduced the percentage of modulation, consequently overmodulation was absent.. The second effect is the introduction of distortion as can be seen by comparison with Figs. 18-16A and C. (G) Trapezoidal trace corresponding to Fig. 18-16F. The trapezoid does not indicate the distortion evident in the wave envelope patterns. (H) Oscillogram of audio output at modulator as the consequence of the impedance mismatch. Compare with Fig. 18-16E. It is evident that improper mismatch is productive of distortion, which is more severe when the load on the modulator is less than required; also that overmodulation is possible when the load impedance is higher than is required.

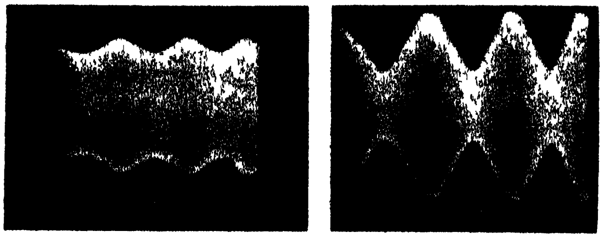


Fig. 18-16.—(I) Oscillogram of modulated wave showing distortion due to mismatch between modulator and r-f final stage at low level of modulation. (J) Increase in percentage of modulation (Fig. 18-16I) results in increased distortion of the modulated-wave envelope.

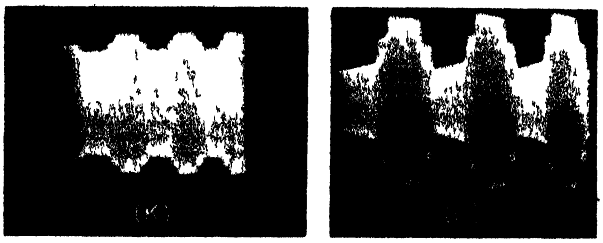


Fig. 18-16—(K) Very bad mismatch modifies pattern of Fig. 18-16I to appear as pattern shown here. Note the low percentage of modulation. The input to the speech amplifier is a sine wave. (L) Pattern indicates wave envelope of carrier modulated at a somewhat higher percentage than in Fig. 18-16J and with equally bad impedance mismatch.

(IMPERFECT NEUTRALIZATION—OSCILLATION)

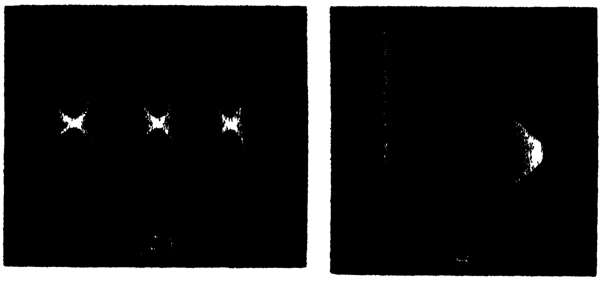


Fig. 18-17.—(A) Oscillogram of modulated wave illustrating the development of oscillation in the r-f final stage at the peak of upward modulation. (Compare with Fig. 18-16A.) (B) Trapezoidal pattern corresponding to Fig. 18-17A. Note steep rise at extreme left which corresponds to peak amplitude of modulated wave.

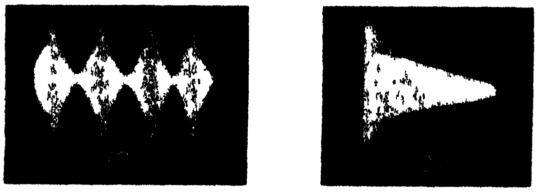


Fig. 18-17—(C) Another example of a modulated wave obtained from a final r-f stage which has been deneutralized. Note that oscillations begin at the start of the upward modulation and cease as the peak modulation is reached. This is sufficient to destroy all intelligibility. (D) Trapezoidal pattern corresponding to Fig. 18-17C.

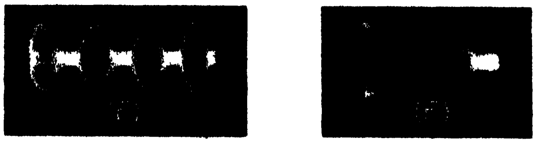


Fig. 18-17.—(E) Wave envelope of modulated carrier obtained from r-f final stage which has been improperly neutralized. Severe distortion results although oscillation in stage is not as evident as in Figs. 18-17 A and C. (F) Trapezoidal pattern corresponding to Fig. 18-17E.

(R-F DETUNING AND IMPERFECT NEUTRALIZATION)

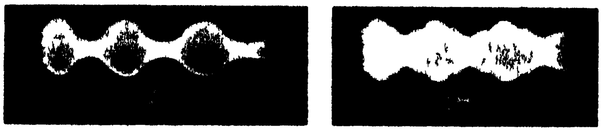


Fig. 18-18.—(A) Another example of the effect of imperfect neutralization of the final r-f amplifier. The distortion is severe. Compare with Fig. 18-17E. (B) This pattern corresponds to Fig. 18-18A except that the plate voltage applied to the r-f tube was reduced, although everything else was kept as before. The distortion is very severe.

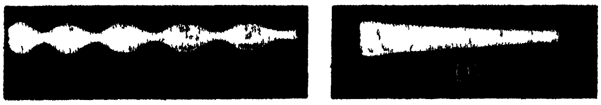


Fig. 18-18.—(C) Oscillogram of modulated wave obtained from push-push *doubling* r-f final when stage is slightly detuned. The presence of three envelopes of like frequency but different amplitude is unaccounted for. (D) Trapezoidal pattern corresponding to Fig. 18-18C.

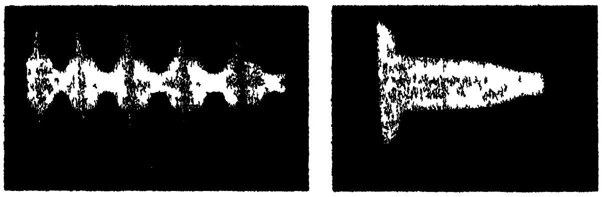


Fig. 18-18.—(E) The stage depicted in Fig. 18-18C (doubling final) detuned. Oscillations developed at the peak of modulation. The distortion of the wave is extremely severe. (F) Trapezoidal trace corresponding to Fig. 18-18E.

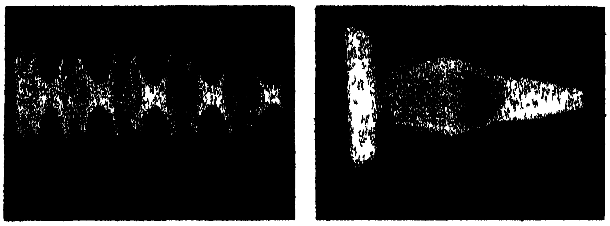


Fig. 18-18.—(G) Another example of oscillation accompanying detuning in a push-push doubling final r-f stage. Oscillation at the peak of modulation can be seen clearly; the presence of harmonics of the driving voltage also are present. (H) Trapezoidal trace corresponding to the modulated wave appearing in Fig. 18-18G.

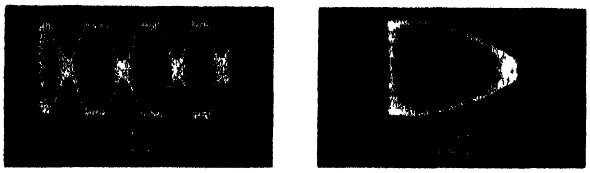


Fig. 18-18.—(I) This is an example of detuning in a straightforward r-f final stage plate tank which has been properly neutralized. Nonlinear operation prevails over most of the upward modulation, whereas substantially linear modulation prevails during the downward modulation. The distortion is very severe. (J) The trapezoidal trace corresponding to Fig. 18-18I. The nonlinearity over a portion of the modulation process is clearly seen.

(MODULATION WITH DISTORTED AUDIO)

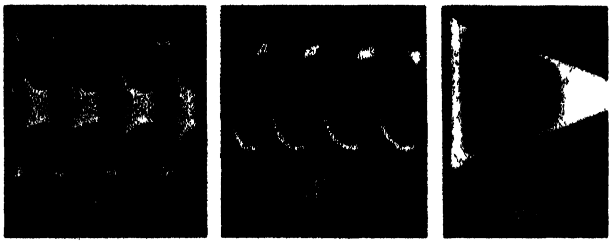


Fig. 18-19.—(A) Oscillogram of modulated-carrier-wave envelope when audio signal is distorted as shown in (B). The audio distortion is caused by overdriven stage in speech amplifier. Modulator system incapable of modulating r-f final. (C) Trapezoidal pattern corresponding to wave envelope in Fig. 18-19A. It is interesting to note the sides of the trapezoid when the r-f and modulating voltages are properly phased at the oscilloscope and when phase displacement exists at the display device. The sides are symmetrical although the modulating signal is distorted. (See Fig. 18-19E.)

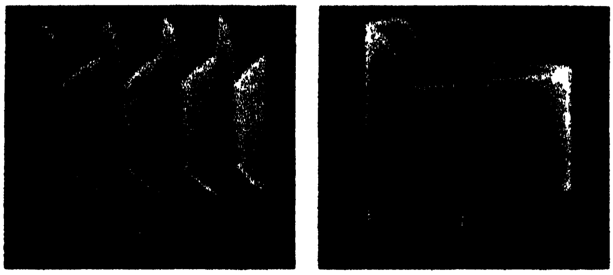


Fig. 18-19.—(D) Oscillogram of modulated-carrier-wave envelope with overdriven speech amplifier. Percentage of modulation comparatively low. Character of envelope indicates another example of distorted audio waveform. (E) Oscillogram of trapezoidal pattern corresponding to Fig. 18-19D. Phase shift between r.f. and audio exists at oscilloscope relative to phase relation at modulator output. Note the correspondence between shape of carrier envelope for forward trace of trapezoid and for the return trace, that is, the steps in the modulation pattern and the trapezoid. (Compare with Fig. 18-19C.) In the absence of phase shift between r.f. and audio at the oscilloscope, the sides of the trapezoid (Fig. 18-19E) would be symmetrical.

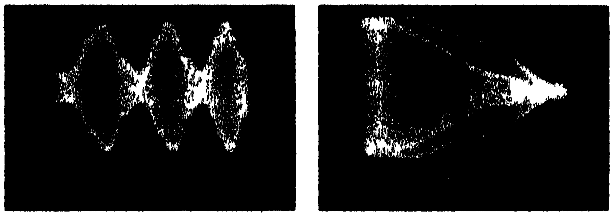


Fig. 18-19.—(F) Another example of modulation when the audio is distorted. The first stage in the speech amplifier was overdriven. (G) The trapezoidal trace corresponding to Fig. 18-19F with phase shift. The kinks in the audio wave are evident in the trapezoid. Obviously the modulating capability of the audio system is inadequate.

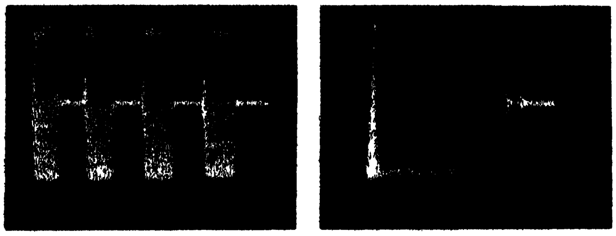
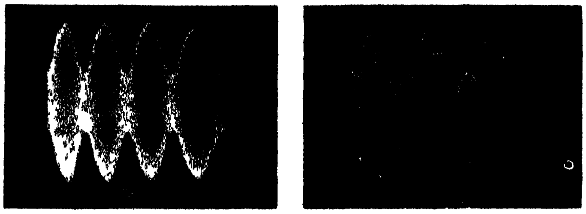


Fig. 18-19.—(H) Oscillogram of overmodulated wave with severe distortion amounting almost to square-wave modulation. (I) Trapezoidal trace corresponding to Fig. 18-19H with phase shift between r.f. and audio at oscilloscope. Shape of audio signal is evident in trapezoidal trace.

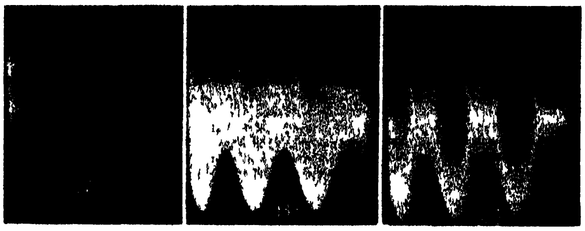
Fig. 18-19.—(J) Another example of severe audio distortion as it appears in the oscillogram of the modulated-wave envelope. (K) Oscillogram of audio output from overdriven modulator stage. (L) Oscillogram of trapezoid corresponding to Fig. 18-19J. Phase relation between r.f. and audio at oscilloscope corresponds to that at modulator. Compare with Fig. 18-19I. The absence of phase shift develops symmetrical sides in the trapezoid although the audio voltage approximates a square wave such as that which caused Fig. 18-19H.

Fig. 18-19.—(M) Oscillogram of modulated wave with unbalanced modulator stage. One modulator tube only partially effective. (N) Oscillogram of modulated wave when modulator stage is overdriven. Compare with Fig. 18-19F. Obviously similar types of distortion can develop at different points in the audio system.



(HUM PATTERNS)

Fig. 18-20.—(A) Oscillogram of hum modulation of carrier. Carrier was modulated at 1000 cycles. (B) Oscillogram of modulated wave with hum present in modulator power supply, filter capacitor open. (C) Oscillogram of modulated wave with hum present in r-f power supply, filter capacitor open. Compare with Fig. 18-20B. Hum in modulator system causes much more complex pattern.



(IMPROPER TEST PROCEDURES)

Fig. 18-20.—(D) Oscillogram of modulated wave (sine-wave modulation) and 60-cycle hum pickup with audio source ungrounded and unshielded connecting leads between devices. (E) Spurious oscillations due to lead dress of unshielded test leads connected between audio modulating source (no modulating signal) and transmitter and oscilloscope. Pattern disappeared when shielded leads were substituted.

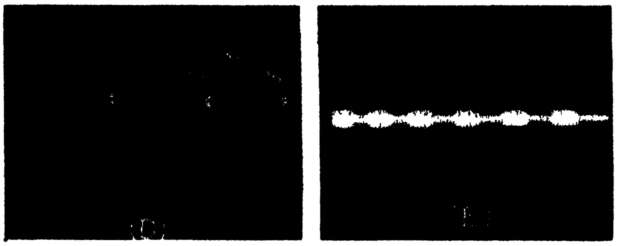


Fig. 18-20.—(F) Distortion of modulated-wave envelope due to overloading in vertical amplifier of oscilloscope when carrier frequency necessitated use of oscilloscope vertical amplifier. (G) Virtual rectification of modulated carrier of Fig. 18-20F when signal level fed into vertical amplifier of oscilloscope was increased.

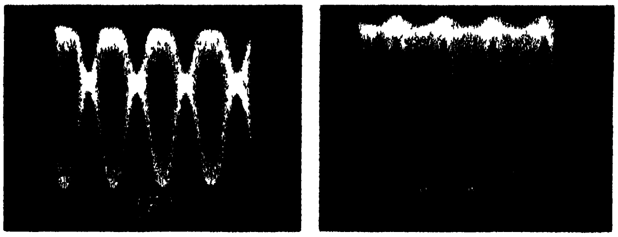


Fig. 18-20.—(H) Oscillogram of modulated-wave envelope under proper test conditions. (I) Illustrates leakage between the input terminals of the oscilloscope's horizontal amplifier and the horizontal amplifier system (which was disconnected) when the linear sweep within the oscilloscope was used to develop the modulated-wave envelope. The input terminals of the horizontal amplifier remained connected to the output of the modulator. Removing the leads from the horizontal amplifier input terminals produced the correct trace as shown in Fig. 18-20H.

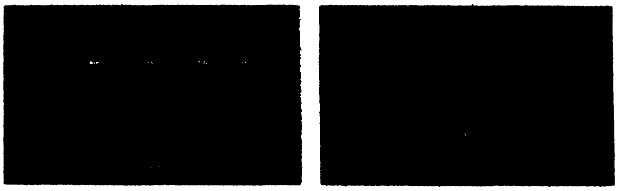


Fig. 18-20.—(J), (K) Another example of the conditions shown in Figs. 18-20H and I, respectively, except that the modulating signal is distorted and modulation is not linear.

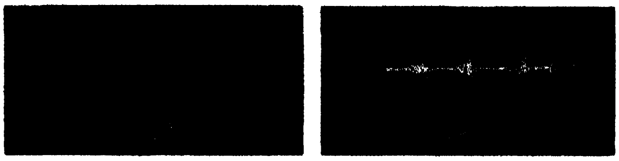
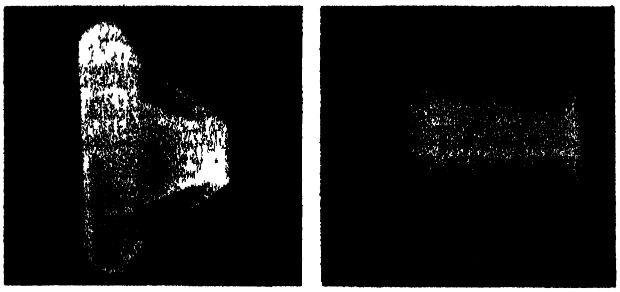


Fig. 18-20.—(L) Oscillogram of trapezoidal trace caused by hand capacity at the leads which fed the modulating signal to the horizontal-deflection plates. With hand capacity removed the pattern became normal and illustrative of proper modulation. (This is not a constant effect, but it may be experienced.) (M) Oscillogram of modulated r-f carrier with linear sweep in oscilloscope adjusted to very low frequency relative to modulating frequency, but with the synchronizing control advanced too far. Compare with other r-f block patterns.



Typical Improper Conditions of Operation

In Figs. 18-15 through 18-20 are shown a number of oscillograms representative of transmitter output when the conditions of operation in the final r-f stage (Class C) and the modulator stage are incorrect. Both modulated-wave envelope and trapezoidal types of patterns are included. The latter more clearly indicate the condition and it is suggested that they be used in preference to the wave envelope variety. It may seem that a number of different conditions are productive of similar effects. This is correct and it is important to realize that it is very difficult at times to select individual contributing causes as being responsible for a certain type of pattern because of the interdependence between operating parameters in the final r-f stage. This should cause no concern, however, because the locale of the improper operation is indicated even when it involves both the modulated and the modulator stages. It then becomes a matter of checking the operating voltages and signal conditions in these stages.

Duplication of the reference patterns demands identical conditions. These are unlikely, therefore, the patterns shown should be accepted as patterns *typical* of conditions. It is likely that tests made on transmitters will result in oscilloscope displays which resemble the reference pattern, and in this connection, the trapezoid trace will most closely approximate the appearance of the reference pattern. This has been our experience over many years of such tests.

Amplitude Modulation by Phase-Shifting Method[2]

There is an interesting method of obtaining amplitude modulation by using two r-f signals of the same amplitude and frequency, varying the phase between them by the audio signal, and combining the two outputs.

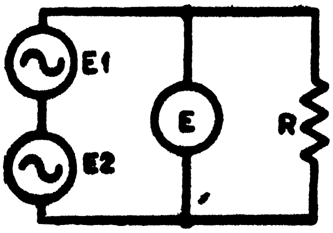
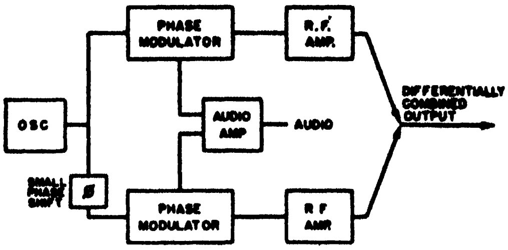


Fig. 18-21.—Series connection of a-c generators used to explain amplitude modulation by phase shifting.

Courtesy CQ

The basic theory may be explained by supposing two a-c generators to be connected in series across a load resistor R as in Fig. 18-21 and considering what happens in the circuit. It is assumed that the generators have the same frequency and voltage, but the phase relation is alterable at will.



Courtesy CQ

Fig. 18-22.—Block diagram of phase-modulation system.

Fig. 18-22 illustrates a setup that may be used to attain the desired objectives. The output of the oscillator is split between two branches of separate amplifiers. The r.f. is brought to the desired power level and is then differentially combined in the output circuit. A phase difference of about 30° is established between the two branches, to provide an unmodulated carrier. When modulation is applied, the phase varies ±30° for 100 per cent modulation. A tube circuit in each line-up is selected to operate as a phase-shifting amplifier, so designed that, upon application of audio, its output will be phase-modulated by a small amount. Where frequency multiplication is to take place further on in the line-up, the amount of the required phase shift is reduced by the factor of the multiplication.

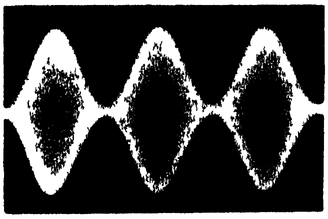


Fig. 18-23.—Oscilloscope pattern obtained when the system is properly adjusted.

Courtesy CQ

An oscilloscope is extremely useful for observing phase shifts between the output components and for recognizing any parasitic oscillations which might occur on modulation peaks. Fig. 18-23 shows the oscilloscope pattern obtained when the system is properly adjusted. The percentage of modulation is nearly 100 per cent. Fig. 18-24 shows the pattern obtained for an overmodulation condition when two amplifier branches contribute unequal outputs. When the two r-f signals are directly opposed (180°), the resultant is a minimum which is not zero; as overmodulation occurs, making the angle greater than 180°, the resultant again increases, giving the reversed negative peak.

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[2] Hartman, W. Herbert, "Neglected outphasing system of modulation," *CQ*, vol. 5, October, 1949. The following material is an abridged excerpt from this paper.

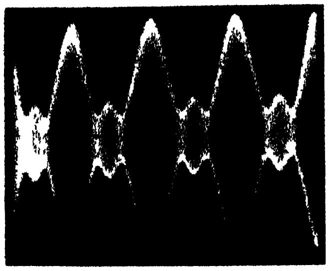


Fig. 18-24.—Pattern obtained for an overmodulation condition when two amplifier branches contribute unequal outputs.

Courtesy CQ

Fig. 18-25 is an oscilloscope pattern that indicates the modulated output when the system is properly adjusted but the audio level is greater than that required for 100 per cent modulation. Although severely distorted after detection in a receiver, this wave contains no sidebands other than the two produced by the sine-wave audio tone at lower percentages of modulation. The sideband relation in this case is that the sum of the two sideband components exceeds that of the carrier, whereas for 100 per cent modulation, their sum is equal to the carrier.

Fig. 18-25.—Modulated output when the system is properly adjusted but the audio level is greater than that required for 100 per cent modulation.

Courtesy CQ

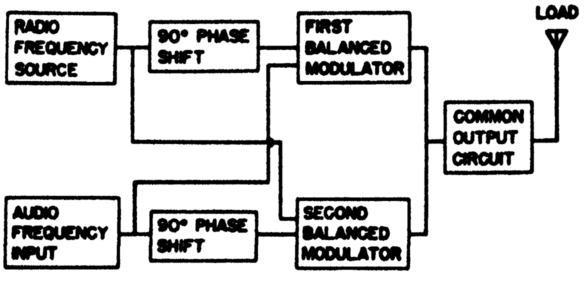
Single-Sideband Operation[3, 4]

In recent years, the single-sideband suppressed-carrier type of transmission has received favorable attention. A phasing system consisting essentially of two balanced modulators whose outputs are combined, operating 90° out of phase with respect to r.f. and a.f., may be used for producing this type of signal.

What the circuit amounts to is two pairs of output tubes, each pair being balanced to provide a double-sideband suppressed-carrier signal, and the two pairs balanced against each other, to cancel one sideband, giving the desired output. A block diagram is given in Fig. 18-26.

If the circuit is not properly adjusted, the carrier will reappear along with any sideband output. Suppression of the undesired sideband, in any sideband cancellation system, is dependent on the accuracy of setting of both the r-f and the a-f phase adjustment. The presence of any sizeable amount of the undesired sideband shows up as a distortion of the envelope of the combined output when the audio input is a sine wave. The undesired frequency component in the output combines with the desired sideband and causes the envelope of the combination to have a "modulation," or ripple, whose frequency is twice that of the audio modulating frequency.

Small inaccuracies in either magnitude or phase setting (a-f or r-f) produce ripples nearly identical in appearance, since, in each case, the effect of the misadjustment is to permit some undesired sideband to appear as a result of incomplete cancellation. This ripple may be conveniently used to find the correct r-f phase setting. Assuming the audio phase shift to be correct, the operator must first equalize the individual balanced modulator output amplitudes, which may be done by adjusting the relative gain of the two a-f input channels. One balanced modulator should be operated at a time, and its output may be conveniently measured with the oscilloscope. The two are set to produce r-f outputs of equal magnitude for a given common modulating signal.



Courtesy QST

Fig. 18-26.—Block diagram of single-sideband system.

The r-f phase may then be adjusted until the ripple on the output envelope disappears. A 10° inaccuracy in phase setting will result in an 8 per cent ripple "modulation" of the single-sideband envelope, which is quite readily visible on an oscilloscope. When highest accuracy of setting is desired, it is best to "touch up" the grid tuning (i.e., the r-f phase setting) until the ripple is at a minimum, while observing the output envelope on an oscilloscope.

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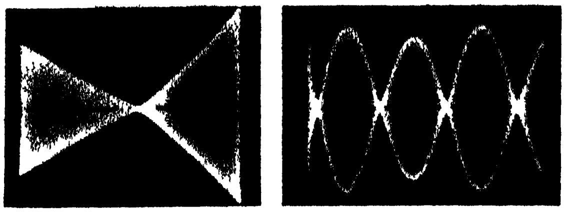
[3] Dawley, Ray L., "An S.S.S.C. transmitter adapter," *QST*, vol. 32, pp. 40-49, July, 1948; the following material contains excerpts from this article.

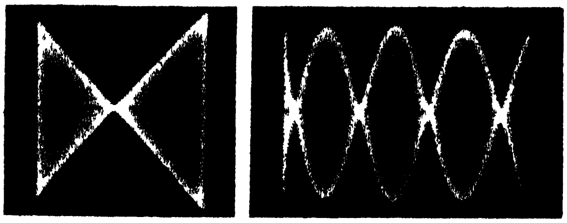
[4] Villard, Oswald G., "A high-level single-sideband transmitter," *Proc. I.R.E.*, vol. 36, pp. 1419-1425, November, 1948; the following material contains excerpts from this article.



Courtesy Proc. I.R.E.

Fig. 18-27.—Lack of equalization is indicated by these oscilloscope patterns.

A similar procedure can also be employed to equalize the outputs of the two tubes in each pair of balanced modulators. One possible way to adjust their outputs is to vary the grid-bias resistors until each tube produces the same r-f output for a given positive d-c screen voltage. To set the resistors, it is best to turn off one balanced modulator. If the remaining pair of tubes is not equalized, the oscilloscope patterns will resemble Fig. 18-27. Adjustment of the grid resistors for equal tube efficiencies will give patterns like those of Fig. 18-28. The process is then repeated for the second pair of balanced modulators. The correct adjustment is that which makes the slopes of the diagonal lines the same in each half of the trapezoidal pattern.



Courtesy Proc. I.R.E.

Fig. 18-28.—Adjustment of the grid resistors for equal tube efficiencies results in these patterns.

In Fig. 18-29 the resistors are set correctly, but the negative screen bias is too great for the applied tube plate voltage and r-f excitation. Both tubes are cut off during a portion of the audio cycle. Fig. 18-28 shows the correct adjustment. (The right-hand half of this pattern is somewhat larger than the left because of stray carrier voltage being picked up from a buffer stage.)

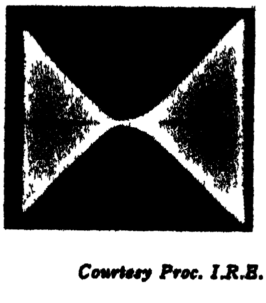


Fig. 18-29.—Example of correct resistor settings but negative screen bias is too great for the applied tube plate-voltage and r-f excitation.

Courtesy Proc. I.R.E.

Fig. 18-30 shows the envelope with both balanced modulators running, but with the undesired sideband not completely cancelled due to inaccuracies in either phase or amplitude setting, or both. The correct phase and amplitude settings produce an output envelope very nearly free from ripple, as shown in Fig. 18-31. The undesired sideband rejection obtainable with the twin balanced modulator circuit can be made as complete as desired by proper adjustment of phase and amplitude settings.

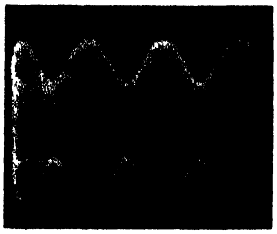


Fig. 18-30. — Envelope pattern obtained with both balanced modulators running, but with the undesired sideband not completely cancelled due to inaccuracies in either phase or amplitude setting or both.

Courtesy Proc. I.R E.

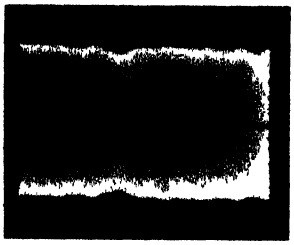


Fig. 18-31.—Correct phase and amplitude settings produce an output envelope very nearly free from ripple.

Courtesy Proc. I.R.E.

A-M Broadcast Tests[5]

The impedance of an antenna, and particularly the common-point impedance of an array, often varies widely over the transmission band. When this is the case, the frequency response, amplitude linearity, and modulation capability of the broadcast transmitter can be adversely affected.

The form of the modulation envelope, seen on the oscilloscope, is greatly different at different points in a coupling circuit or at different points along a transmission line when the impedance of the termination differs substantially from the impedance at the carrier

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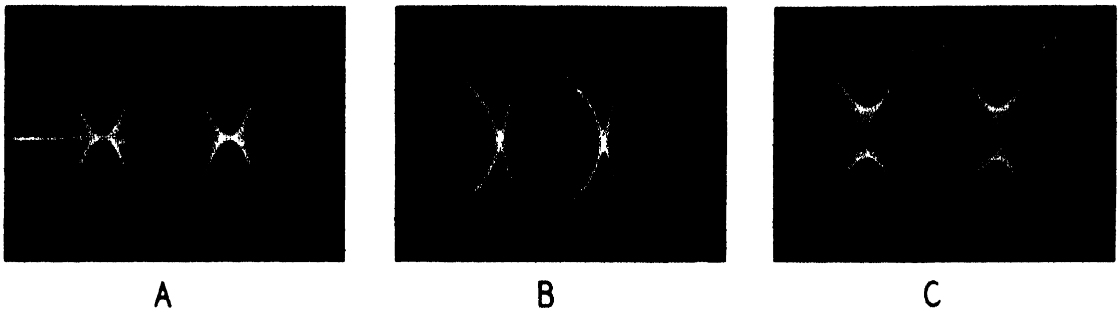
[5] Doherty, W. H., "Operation of AM broadcast transmitters into sharply tuned antenna systems," *Proc. I.R.E.*, vol. 37, pp. 729-734, July, 1949; the following material contains excerpts from this article.



Courtesy Proc. I.R E.

Fig. 18-32.—Oscillograms which show the voltage envelope for a modulation frequency of 7,500 cps as observed at three points along a transmission line or coupling network; (A) fully modulated voltage wave, (B) distorted voltage wave at odd eighth-wavelength points, (C) wave at odd-number quarter-wavelength points.

frequency. This is illustrated by the oscillograms of Fig. 18-32, which show the voltage envelope for a modulation frequency of 7,500 cps as observed at three points along a transmission line or coupling network whose termination, for example, is equivalent to a series-tuned circuit resonant at the carrier frequency. At the termination or at points removed therefrom by an even number of quarter-wavelengths, a fully modulated voltage wave may appear, Fig. 18-32A, and the amplitude of each of the two side-frequency *voltages* accordingly will be one-half the amplitude of the carrier frequency; the side-frequency *currents* will each be less than one-half the carrier current, and an inspection of the current envelope would show substantially less than 100 per cent modulation.

On the other hand, at points removed from the termination by an odd number of quarter-wavelengths, the impedance will correspond to that of a parallel-tuned circuit, decreasing on either side of the carrier frequency; although the current envelope, if observed, would show a fully modulated wave, the voltage envelope, Fig. 18-32C, shows considerably less than 100 per cent modulation, in the case illustrated, only 60 per cent of the impedance at the carrier frequency.

Finally at odd eighth-wavelength points, where the impedance versus frequency curve is dissymmetrical, the voltage envelope has the distorted appearance indicated in Fig. 18-32 B. The current envelope at this point would also be badly distorted. A monitoring rectifier and conventional distortion-measuring instrument would register a high percentage of distortion for this wave, yet there are no extraneous side frequencies being radiated, the envelope distortion being entirely due to the inequality and phase dissymmetry of the two desired side-frequency voltages at this particular point in the line or coupling circuit.

It is obvious that, if the operator were to raise the audio input level, endeavoring to bring about apparent full modulation at this point, much more severe distortion would be registered, because the wave at other points would then be overmodulated.

Effective Modulation and Distortion

If one wishes to determine actual delivered sideband power by measurement of a sample of the current envelope, it is necessary to make the inspection and measurement at a point in the circuit where the series resistance is independent of frequency over the band transmitted, since it is the current squared times the series resistance that determines power.

In order that the measurement may include true radiated distortion power, i.e., power in extraneous sidebands, the constancy should hold over a correspondingly wider band. Now, when measuring at such points, one should not try to adjust the audio input for 100 per cent modulation of the current envelope observed on the oscilloscope, but for a certain lower percentage, 60 per cent in the case considered in Fig. 18-32, since the voltage envelope (not being observed by the operator but shown in Fig. 18-32A will then have reached full modulation, and any further increase would necessarily bring about severe distortion in both the voltage wave and the current waves.

If, on the other hand, it is the voltage envelope rather than the current envelope that is to be monitored, the point of measurement should be one where the *parallel* rather than the series resistance is constant over the transmitted band, since the power is the square of the voltage divided by the parallel resistance.

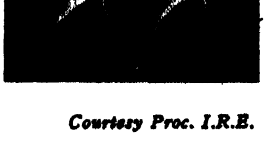
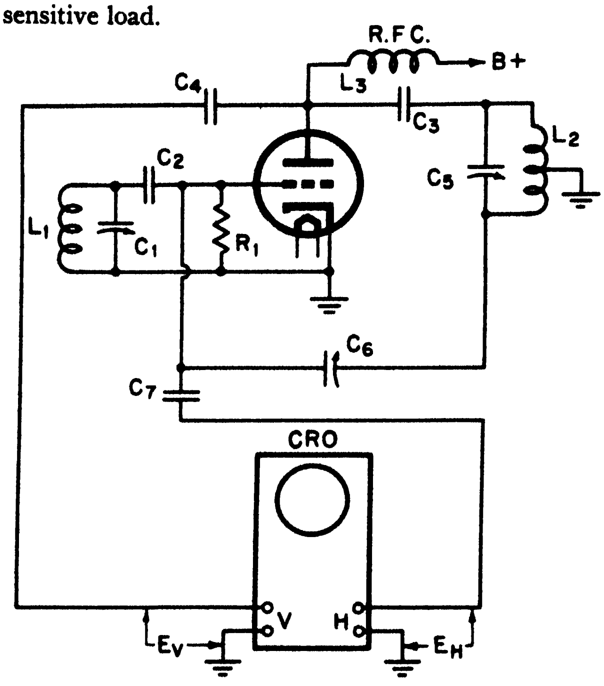


Fig. 18-33.—Modulation envelope of a 10-kw transmitter for a frequency-sensitive load.

Courtesy Proc. I.R.E.

Power Amplifier Operation

With the impedance versus frequency characteristic varying widely from point to point in a coupling circuit, it is unnecessary to state that the characteristic found at the particular point where the power amplifier tubes are connected is of importance in the performance of the amplifier, regardless of the type of circuit or modulation used. Fig. 18-33 shows the modulation envelope of a 10-kw transmitter for a frequency-sensitive load.



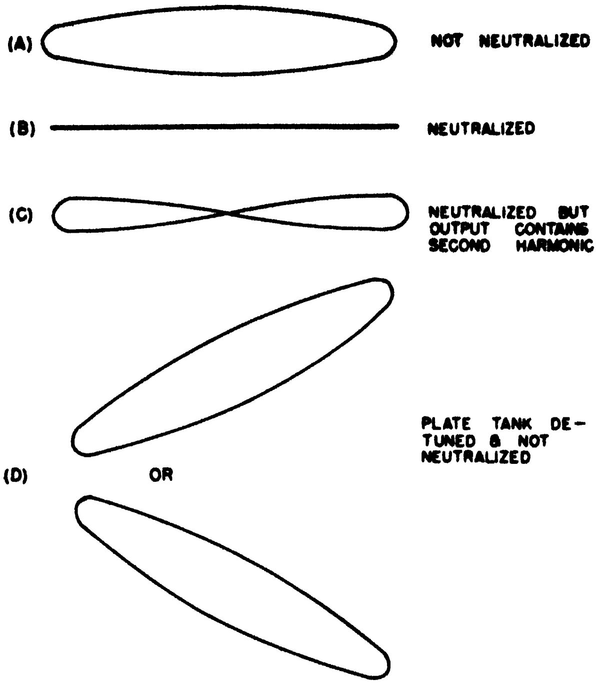
Courtesy Du Mont Labs.

Fig. 18-34.—Oscilloscope arrangement used for checking neutralization of an r-f stage in a transmitter.

Neutralization

In Fig.18-34 the neutralization setup is shown. *Without d-c plate voltage,* r-f voltage is applied through *C4* to the vertical input of the oscilloscope, and the grid voltage through *C7* to the horizontal input. The neutralization procedure may be specified in the following steps:

1. Tune the grid tank circuit *L1-C1* for maximum horizontal-deflection voltage E_H.
2. Tune the plate tank circuit *L2-C5* to produce a horizontal ellipse, as shown in Fig. 18-35 A.
3. Tune *C6,* the neutralizing capacitor, to obtain a straight horizontal line as shown in Fig. 18-35B. The patterns shown in Figs. 18-35C and 18-35D may also be observed under the indicated conditions.


(A)
NOT NEUTRALIZED
(B)
NEUTRALIZED
(C)
NEUTRALIZED BUT OUTPUT CONTAINS SECOND HARMONIC
PLATE TANK DE-TUNED & NOT NEUTRALIZED
(D)
OR

Courtesy Du Mont Labs.

Fig. 18-35.—Oscilloscope patterns obtained for various conditions of correct and incorrect neutralization adjustments.

Measuring Amplitude Modulation at the Receiver

A method of measuring the percentage of modulation of an a-m signal, which is of interest but somewhat limited in accuracy, makes use of a simple principle; the development of a phase difference between the vertical- and horizontal-deflection voltages of the oscilloscope.

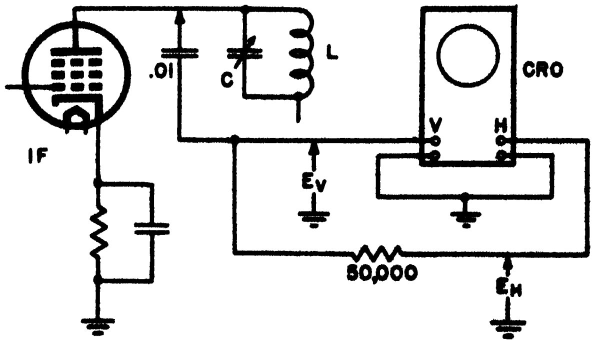


Fig. 18-36.—Circuit for measuring amplitude modulation.

The basic circuit is shown in Fig. 18-36. In Fig. 18-37 sketches illustrating the patterns are given, and the formula for determining modulation percentage. The unmodulated pattern is an ellipse having a single, sharp line. This line broadens to a ribbon with modulation. For 100 per cent modulation, the dark area in the center of the pattern decreases to zero. This point may be determined with accuracy.

The effect of overmodulation is to go beyond this point and a bright spot appears in the center of the pattern. The method is limited in accuracy because of the effect on the signal of the receiver tuned circuits and stages, in any one of which nonlinearity will be a factor. It is also limited by transmission conditions. Noises, heterodynes, interference, etc., will affect the signal.

The coupling to the vertical input of the oscilloscope should be as loose as possible to minimize loading of the plate circuit of the final i-f stage, and retuning probably will be needed. The shift in phase is caused by the horizontal amplifier input capacitance and the 50,000-ohm resistor.

Thus, the voltages E_v and E_H are out of phase, and the out-of-phase voltage relationships result in the development of the patterns shown.

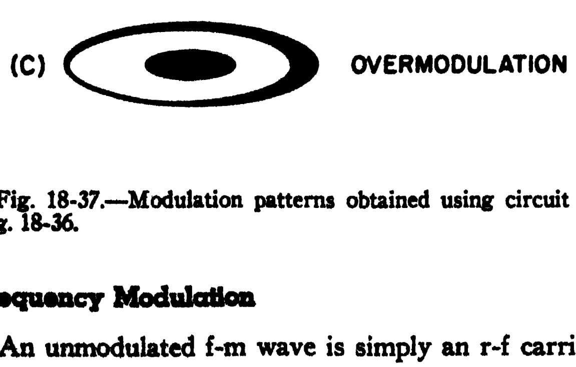


Fig. 18-37.—Modulation patterns obtained using circuit of Fig. 18-36.

Frequency Modulation

An unmodulated f-m wave is simply an r-f carrier wave, which may be shown as a block pattern on the tube screen, as previously indicated. When the wave is frequency-modulated, however, conditions are fundamentally different and auxiliary equipment must be used in conjunction with the oscilloscope for the display of f-m modulation.

Modulation Considerations

In frequency or phase modulation, there is no condition exactly equivalent to percentage of amplitude modulation or overmodulation. A definition of percentage of modulation must be somewhat different for f.m. or p.m. From a practical viewpoint, 100 per cent modulation is obtained when the transmitted signal channel width is equal to the receiver bandwidth. This bandwidth is determined by the design and alignment of the receiver tuned circuits. With wider frequency swing, exceeding the receiver bandwidth results in distortion. The distortion produced in overmodulation of a.m. would be somewhat similar.

To analyze the output of an f-m transmitter, it is first necessary to convert the f-m variations to amplitude changes and then to rectify the r-f voltages. The resulting audio component may then be fed to the input of an oscilloscope. Other means, such as the "panoramic" system, discussed at the end of this chapter, may also be used for spectrum analysis of the frequency-modulated wave.

There are sidebands produced by frequency or phase modulation of a wave, as well as by amplitude modulation. The sidebands are, however, quite different. In the case of f.m. and p.m., single-tone modulation produces a series of pairs of sidebands that bear a harmonic relationship to the modulating frequency. In a.m., of course, there is only one pair of sidebands.

The relationship between the modulating frequency and the carrier frequency deviation determines the number of sidebands that occur in f.m. and p.m. The modulation index is defined as the ratio between the frequency deviation in cycles per second and the modulating frequency in cycles per second. That is:

$$\text{Modulation Index} = \frac{\text{Carrier Frequency Deviation}}{\text{Modulating Frequency}}$$

The modulation index may also be defined as the phase shift in radians. The index is invariable in p.m. regardless of the modulating frequency; in f.m. the modulation index varies as a function of the modulating frequency. Further information on f.m. is given in standard references.[6]

F-M Analyzer

An instantaneous picture of the signal components of an f-m wave may be obtained with a suitable oscilloscope setup, shown in Fig. 18-38 By calibrating the screen in terms of frequency deviations, the frequencies of individual sidebands can be determined.

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[6] Rider, J. F., and Uslan, S. D., "FM Transmission and Reception," John F. Rider Publisher, Inc., New York, N. Y., 1948.

The equipment consists fundamentally of a superheterodyne which picks up the f-m signal. The center frequency of the receiver oscillator is 2 Mc below the input signal carrier. The sweep voltage is linear to produce a variation of +100 kc in the oscillator frequency. This variation occurs at the rate of thirty times per second.

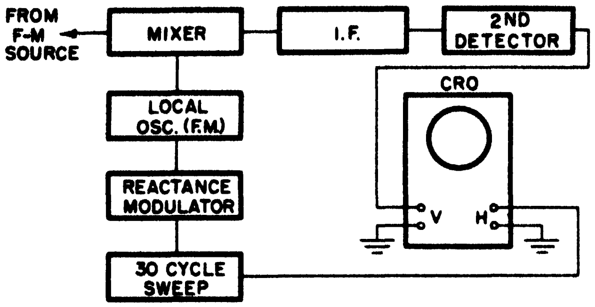


Fig. 18-38.—Setup for obtaining instantaneous picture of the signal components of an f-m wave.

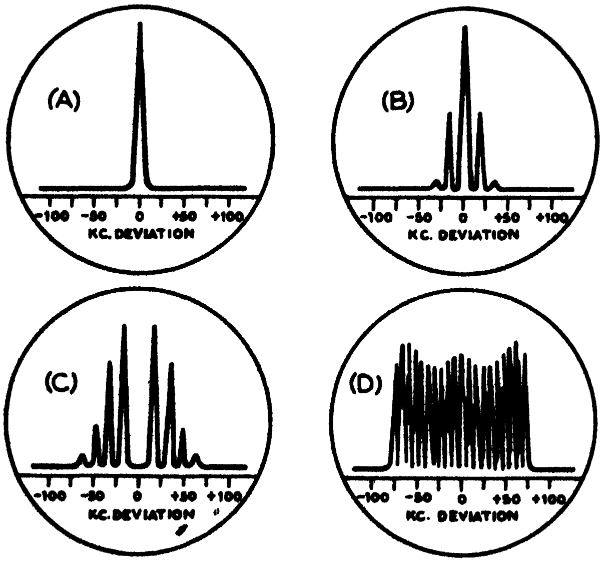


Fig. 18-39.—Representation of pulse deflections on cathode-ray-tube screen in f-m analysis with the oscilloscope.

The result, essentially, is a scanning of the frequency range 100 kc above and below the f.m. signal center frequency. The i-f amplifier responds sharply to any beat at the intermediate frequency produced by the local oscillator and some component of the f-m wave at a particular frequency in the swept band.

The i-f amplifier is sharply tuned to separate the individual components from the complete signal. The second detector of the receiver then demodulates the individual components, since it is an a-m type, and delivers an output whose peak is proportional to the amplitude of the component which is being passed through the i-f system and demodulated.

The various pulses or peaks are distributed on the tube screen as shown in Fig. 18-39. The spacing of the pulses on the screen is set by the 30-cps linear sweep voltage. This voltage is also applied to the reactance tube modulator which frequency modulates the local oscillator in the radio receiver circuit. Fig. 18-39A shows the pattern for an unmodulated f-m carrier, Fig. 18-39B a 7.5-kc deviation with a modulating signal frequency of 15,000 cps, Fig. 18-39C a 36-kc deviation for the same modulating frequency of 15,000 cps and 18-39D a 75-kc deviation for a 2,500 cps audio signal.

Phase Modulation (P.M.)

In a p-m system, frequency modulation necessarily occurs, since the rate at which its phase changes determines the frequency of an alternating current. When the phase changes quickly, the current has a higher frequency than one having a relatively slow rate of phase variation. If the phase variation is 360° per second, the frequency is one cycle per second. For three complete cycles, the phase change in degrees = 3 × 360 = 1,080°.

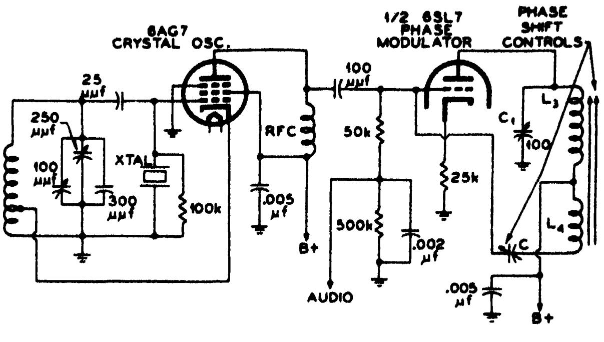
Together with the phase shift, a frequency change occurs; phase advancement corresponds to an increase in frequency; phase retardation is analogous to a decrease of frequency. The rate of phase shift is proportional to the amplitude of the modulating signal and to its frequency and is also proportional to the total number of degrees through which the phase is displaced. In f.m., the frequency deviation varies with the amplitude of the modulating signal but does not change as a function of the modulation frequency. Both the amplitude and frequency, in p.m., produce proportional deviations. It is impossible to distinguish between f.m. and p.m. when there is only one modulating frequency.

Experimental F-M Transmitter Output Waveforms

The display of frequency- or phase-modulated signals is not as simple as it is for amplitude-modulated voltages. Contained herein are a number of oscillograms made during some experiments conducted on a phase-modulated transmitter which was developed for narrow-band f-m operation in the amateur radio bands. The oscillator and modulator circuit of the transmitter is shown in Fig. 18-40. In some respects, this equipment is like transmitters conceived for narrow-band operation in other services and it is more than likely that the conditions shown herein will be paralleled with other apparatus, and the conditions stated may be duplicated in the general run of f-m transmitters provided that similar elements are subject to variation.

Lack of time prevented a more detailed analysis than is contained herein. Perhaps others may pick up where we left off, in which case we certainly would appreciate details for inclusion in subsequent editions. Further analysis will be carried on and we hope to include more data in the revisions of this text.

In view of the frequencies involved, a very stable crystal-controlled converter substantially free from distortion was developed and, where indicated, was used in conjunction with the f-m exciter so as to allow feeding the signal to the vertical-deflection plates through the vertical amplifier. The operating frequency of the converter was selected so as to produce an output frequency about midway between the pass-band limits of the vertical amplifier of the oscilloscope. Oscillograms representative of different conditions are shown in Fig. 18-41. Sometimes the output of the f-m exciter was taken directly from the unit and fed to the vertical plates of the oscilloscope, and more frequently, it was heterodyned to the lower frequency and displayed along a linear time base, using the sawtooth sweep in the oscilloscope. The method which was used is indicated in the captions for the illustrations.



Courtesy Sonar Radio Corp.

Fig. 18-40.—Circuit of oscillator-modulator system of experimental unit.

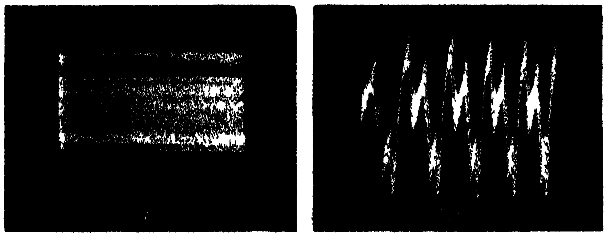


Fig. 18-41.—(A) Oscillogram of unmodulated output of crystal oscillator in f-m exciter. It was link coupled to vertical-deflection plates and the linear sweep was set at a very low frequency. Bands across the trace indicate presence of very strong harmonics. (B) Oscillogram of several cycles of unmodulated carrier corresponding to Fig. 18-41A, developed by use of a converter to beat down carrier to a frequency suitable for application to vertical amplifier of oscilloscope. Time base in the oscilloscope is set to show 5 cycles of unmodulated carrier voltage.

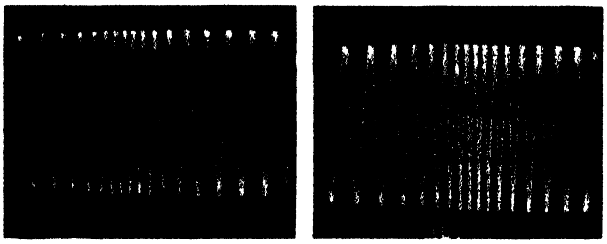


Fig. 18-41.—(C) Sine-wave frequency-modulated carrier after heterodyning by converter and being passed through selective circuit which eliminated harmonics but passed frequency-modulated signal. Approximately 1 audio cycle. All circuits adjusted for normal operation. (D) Same as Fig. 18-41C but at different time.

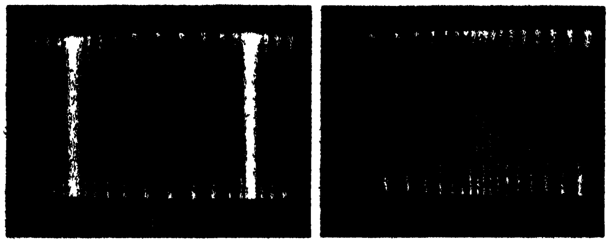


Fig. 18-41.—(E) Like Fig. 18-41C but at different time, and approximately 2 audio cycles modulation. (F) Like Fig. 18-41C but at different time and equal to about 1 audio cycle modulation.

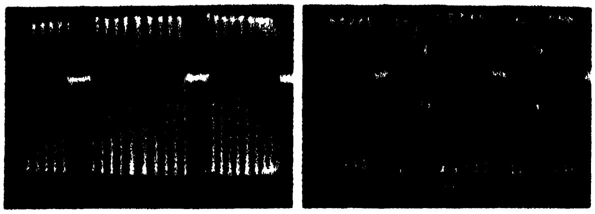


Fig. 18-41.—(G) Oscillogram of sine-wave frequency-modulated signal with incorrect condition of resonance in modulator stage and maximum deviation. Compare with Fig. 18-41C. Note carrier cutoff in Fig. 18-41G at what seems to correspond to the positive peaks of audio. (H) Same conditions as in Fig. 18-41G except for different condition of detuning. Note appearance of amplitude modulation.

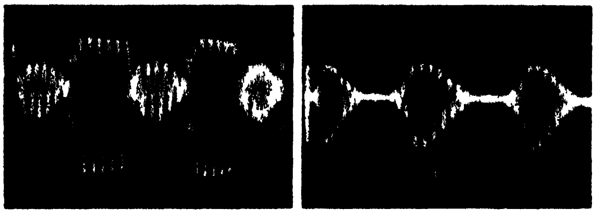


Fig. 18-41.—(I) Same conditions as in Fig. 18-41G except for different condition of detuning. Amplitude modulation is pronounced. (J) Same conditions as in Fig. 18-41G except for different condition of detuning.

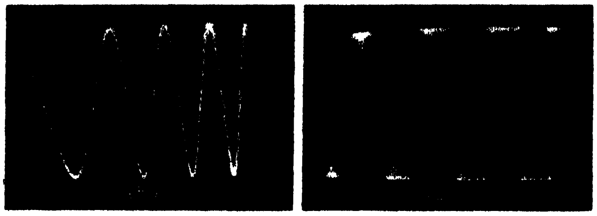


Fig. 18-41.—(K) Oscillogram of f-m wave for part of an audio cycle modulation for comparison with following Figs. 18-41L and M. All conditions normal. (L) Oscillogram of f-m wave shown in Fig. 18-41K but with phase-shift control in the modulator advanced. Note appearance of peaks of wave and compare with Fig. 18-41M.

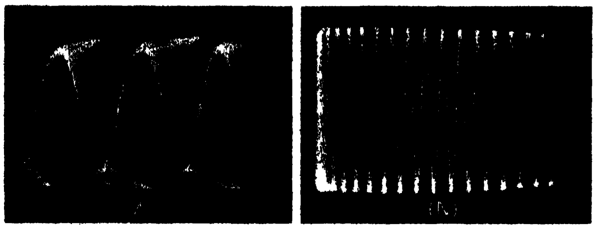


Fig. 18-41.—(M) Oscillogram of f-m wave shown in Fig. 18-41K but with phase-shift control in modulator advanced and plate circuit out of resonance. Note change in peaks of wave. Obviously, condition of resonance affects phase shift. (N) Oscillogram of f-m wave similar to Fig. 18-41C, D, E and F, except with increased deviation. Note spreading of the wave at positive audio peaks in a manner similar to that shown for an increase in phase shift.

TELEVISION TESTS[7]

Certain aspects of television testing remain obscure because of the need for new techniques. However, some tests have been developed, which will be described here.

Testing the Television Transmitter[8]

The frequency characteristic of a video amplifier must be capable of essentially flat response to an upper limit of 5 or 6 Mc. The low frequencies may be tested

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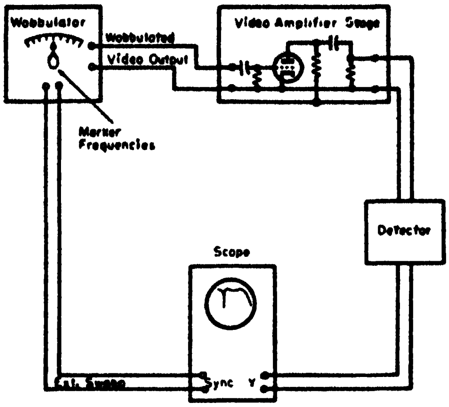
[7] "Standards on Television—Methods of Testing Television Transmitters," The Institute of Radio Engineers, New York, N. Y., 1947.

[8] Hamilton, Edward G., "TV transmitter design," *Communications*, vol. 28, pp. 10-13, July, 1948; the following material contains excerpts from this article.



Courtesy Communications

Fig. 18-42.—Test setup of the wobbulator and video amplifier used to determine frequency response.

best by the application of square waves to the input and observing the amount of tilt on an oscilloscope whose low-frequency characteristic is flat down to at least 10 cps. The tilt allowable in a video amplifier is of the order of two per cent maximum with a 60-cps square wave.

Low-frequency compensation may be used but must not be excessive. The compensation may be observed on the oscilloscope as a rise in the trailing edge while undercompensation may result in a rise of the leading edge. In every case, each stage should be flat by itself. Staggering stages may result in objectionable phase shift and should be avoided wherever possible.

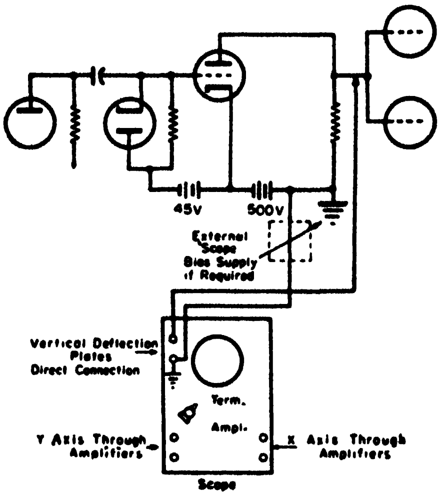
The high-frequency response of video amplifiers may be tested best by means of a beat-frequency "wobbulator" whose output is linear from about 200 kc to 8 Mc, and an oscilloscope. The characteristic of this oscilloscope need not be too exacting since a low-frequency wobbulator sweep is used, and a high-impedance linear detector is employed to measure the amplitude response. Fig. 18-42 shows the test setup of the wobbulator and video amplifier used to determine frequency response.

Testing the D-C Restorer

The function of the d-c restorer is to refer all composite video information to the bias level of the modulator. The restoring characteristic may be measured at two points, namely: (1) at the grid of the modulator, or (2) at the plate of the modulator. Fig. 18-43 shows how an oscilloscope may be connected for this measurement. Since the end result of the restorer is to keep the sync peak at a constant bias reference in the modulated r-f amplifier, it has been found desirable to measure d-c restoration at the plate of the modulator (in effect, at the grid of the modulated amplifier).

The connection to the oscilloscope must be made directly to the deflection plate since the d-c component is the factor in question. With no video signal, the beam on the oscilloscope is adjusted for center position by means of the positioning control. The video signal is applied to the video amplifier and the peak sync position on the screen must remain at essentially the center position, as adjusted above, for proper operation.

In case the beam is deflected off the face of the cathode-ray tube, an external positioning bias is required which will equalize the static potential. It is important to note that this removes the oscilloscope chassis from ground potential; therefore, no direct ground connection should be made where external positioning is required.


45v
500v
External Scope Bias Supply if Required
Vertical Deflection Plates Direct Connection
Term.
Ampl.
Y Axis Through Amplifiers
X Axis Through Amplifiers
Scope

Courtesy Communications

Fig. 18-43.—Circuit for measurement of d-c restorer circuit characteristics.

R-F Pass-band

The r-f pass-band characteristic is measured on the master series transmitter by means of a built-in r-f wobbulator whose frequency range is ±5 Mc about the carrier frequency. Diode pickup sampling circuits are installed in the plate circuit of the modulated amplifier, cathode circuit of the intermediate power amplifier, cathode circuit of the power amplifier, plate circuit of the power amplifier, and the transmission-line output.

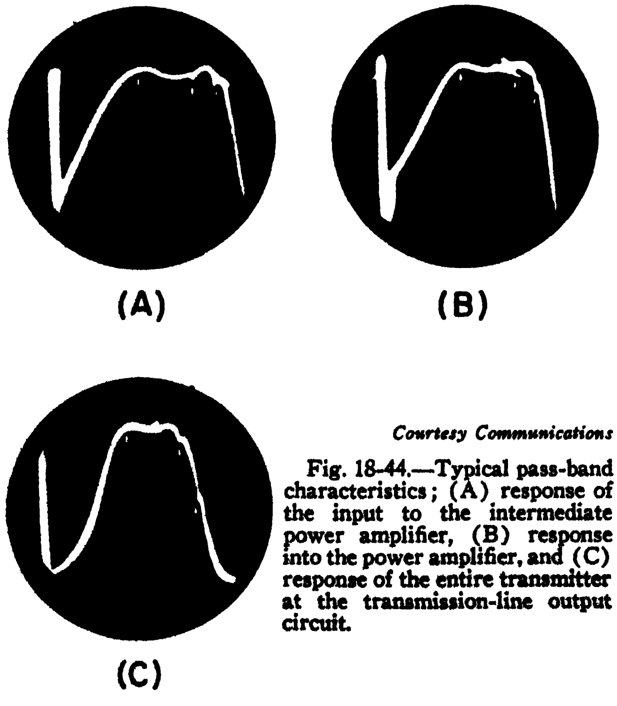
An oscilloscope is provided to monitor any three of these pickup points simultaneously. Marker frequencies are provided by injecting the carrier frequency into the modulated amplifier grid circuit, and by inserting absorption type wave traps, tuned to high and low frequency ends of the pass-band characteristic, in the plate circuit field of the modulated amplifier. The modulated amplifier grids may be connected, by suitable switching arrangements, to either regular r-f drive or wobbulator drive.

Adjustments

After initial adjustment of the r-f stages is accomplished, only a few minutes are required to check performance or compensate for tube replacements, etc. Initial adjustment, however, requires a fundamental approach to the problem as previously described.

The coupling between stages is reduced to a minimum, and gradually increased until a double hump response is obtained with the low-frequency peak to approximately 3 Mc higher than the carrier frequency, these conditions being observed on the wobbulator oscilloscope at the various pickup points.

The cathode loading and coupling determines the frequency separation and flatness of the response characteristic. Frequency separation is predominantly affected by coupling, while flatness is a function of loading. Loading is varied by adjusting the point of the cathode connection. Movement toward the open end of the resonant transmission line results in increased loading. A fine adjustment of this parameter is provided by the movable shorting bar; this factor also has an effect on coupling since the area of coupling is affected by its movement.



Courtesy Communications

Fig. 18-44.—Typical pass-band characteristics; (A) response of the input to the intermediate power amplifier, (B) response into the power amplifier, and (C) response of the entire transmitter at the transmission-line output circuit.

The foregoing procedure applies to the circuits between the modulated amplifier plate and intermediate-power-amplifier cathode, and power-amplifier plate and intermediate-power-amplifier cathode. The output circuit is tuned as indicated above, except that increased loading is accomplished by moving the transmission-line connections toward the open end of the resonant-line output circuit.

Since three oscilloscopes are provided for observing three patterns simultaneously, any interaction between stages may be observed and corrected without the necessity for many retest checks. The diode samplers most useful for these adjustments are located in the intermediate-power-amplifier cathode circuit, and transmission-line output circuit.

Fig. 18 44 shows a typical pass-band characteristic for each of the three stages. The bandwidth was adjusted as shown in the first photograph. Curve *A* is the response of the input to the intermediate power amplifier. Curve *B* is the response into the power amplifier, and curve *C* is the response of the entire transmitter at the transmission-line output circuit.

The low-frequency end of the sweep is at the left side. The marker shown up near the top of the band-pass characteristic represents 77.25 Mc (for channel 5 operation). The second marker to the right represents 80 Mc or a frequency deviation of 2.75 Mc. The third marker to the right represents a frequency of 81.5 Mc or a deviation of 4.25 Mc from the carrier. The vertical line at the left of each curve is of no significance in this discussion. It will be noted that the lower side of the response characteristic becomes steeper as it passes through the amplifiers, resulting in adequate suppression of the lower sideband.

VIDEO SIGNAL MONITORING[9]

To evaluate the quality of the composite video signal used for television transmission, it is necessary to use three units:

(1) A *line* waveform monitor which displays the video waveform on a cathode-ray tube, the horizontal sweep being linear and synchronized at half the line frequency of the video signal, i.e., 7,875 sps (sweeps per second).

(2) A *frame* waveform monitor, similar to (1), but with its sweep frequency equal to half the *field* frequency of the video signal, i.e., 30 sps.

(3) A *picture* monitor capable of high resolution.

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[9] Ruston, John, "TV control console design," *Communications*, vol. 28, pp. 8-11, October, 1948; the following material contains excerpts from this article.

The video inputs of the two waveform monitors are usually connected in parallel, and since their horizontal sweeps are half-line and half-frame frequencies, a line sync pulse and a frame sync pulse appear at the centers of their respective traces. By means of suitable scales in front of the cathode-ray-tube screens, the relative magnitudes of the sync pulses and picture content can be noted, and any irregularities in the composition of the video signal can be observed. The picture content of the video signal cannot be interpreted on the waveform monitors, so the picture monitor is required for this purpose. This shows faults such as lack of definition, ghosts, white saturation, poor contrast, etc.

To check the operation of the visual transmitter, it must be possible to monitor the video input to the transmitter and also the video signal which appears as a modulated r-f voltage on the transmission line to the antenna. It is desirable to monitor intermediate points in the transmitter and also by means of a monitoring receiver which picks up the signal *off the air*. Hence, switching must be provided in the video inputs to the monitors, and since the range of frequencies in the video signals is from 30 cps to 6 Mc, problems arise in this connection which are not present when switching audio circuits.

Bridging Connections

Video signals are usually transmitted in coaxial cables of 75-ohm surge impedance; therefore, if such a line is to be bridged by a monitoring connection, the bridging impedance must be high, compared with 75 ohms, at all frequencies up to 6 Mc.

Two waveform monitors and one picture monitor have to be bridged across the line and their total input capacitance, together with the capacitance of connecting cables and switches, cannot be reduced below about 150 $\mu\mu f$, which has an impedance of about 170 ohms at 6 Mc. The connection of this capacitance across the video input to the transmitter would affect the frequency response and introduce reflections.

By using the potential dividing circuit of Fig. 18-45, the effect of the bridging capacitance can be reduced to a negligible amount. The inductance L and capacitance C are adjusted so that the effective impedance between the point A and ground is almost a pure resistance at frequencies up to 6 Mc, the theory of this operation being similar to the use of peaking coils in the plate circuits of video amplifiers for correcting high-frequency response.

The addition of the 80-ohm resistor provides a correct resistive termination for the 75-ohm video line.

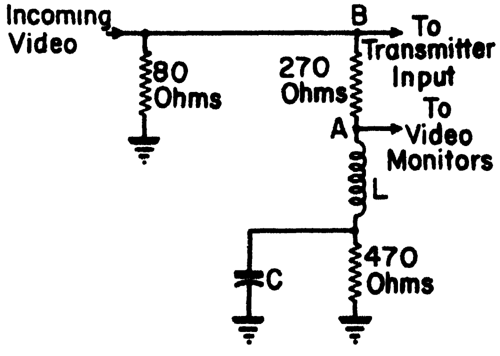
The monitor input switches must be connected to point A, and when a monitor is disconnected, it must be replaced by an equivalent capacitance.

During the instant of switching, assuming that non-shorting type switches are used, the capacitance to ground, from point A, will be reduced momentarily, but this will only have a slight effect on the high-frequency response in the video line and is not noticeable in the picture.

By using this bridging circuit across the line, the monitors can be switched to monitor any video line without having any appreciable effect on the signal in the line. The only requirement is that the monitors and switch shall be located fairly close to the termination of the line to avoid excessive capacitance in the connecting cables.

An alternative method of bridging a video line is to use a bridging amplifier or a cathode follower. The grid of an amplifier tube can be switched to a video line without introducing an appreciable impedance change and the monitors can then be connected to the output of the amplifier.

Any switching surges or impedance mismatch caused when the monitors are switched to the amplifier output are isolated from the video line.

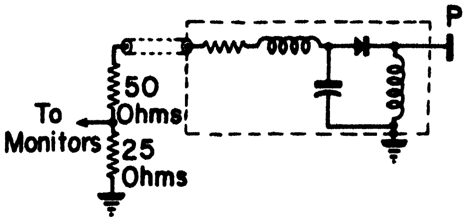


Courtesy Communications

Fig. 18-45.—Potential dividing circuit by means of which the effect of the bridging capacitance can be reduced to a negligible amount.

Terminating Connections

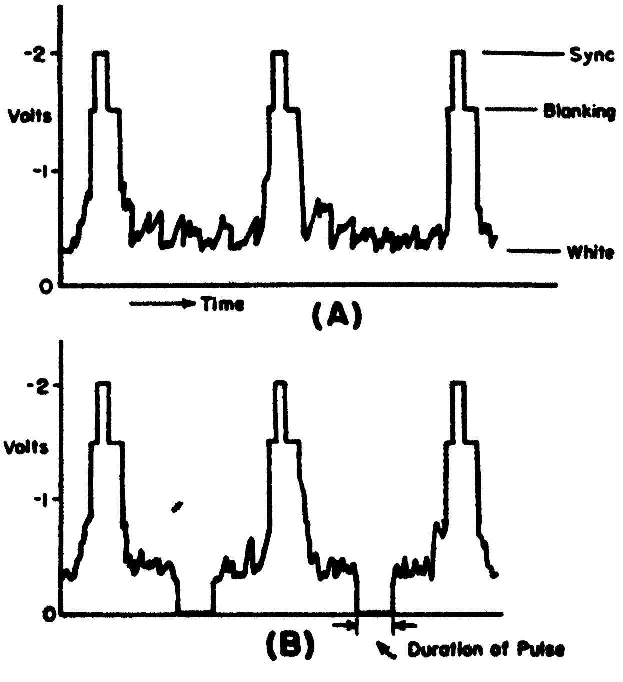
The bridging connections described are used only on the video input to the transmitter or where connections of monitors to a video line must not influence the signal in the line. At other monitoring positions in the transmitter, a video signal is produced solely for supplying the monitors and the coaxial lines conveying these signals to the console must be terminated with 75 ohms.



Courtesy Communications

Fig. 18-46.—Simple detector circuit for obtaining video monitoring signals.

Video monitoring signals can be obtained in the transmitter by using the simple detector circuit shown in Fig. 18 46. The pickup wire *P* is placed close to the r-f circuit to be monitored and, by moving the position of *P*, the amplitude of the video signal can be adjusted to the required peak value of about 2 volts peak to peak. The potential divider shown is used to terminate the video line at the console, the monitor being switched across the 25-ohm section and thus receiving a signal of ⅔ volt peak to peak. The connection of the monitor capacitances across the 25 ohms will not influence the frequency response to the monitor inputs excessively and any switching surges are of no consequence since the video line is connected only to the monitors.



Courtesy Communications

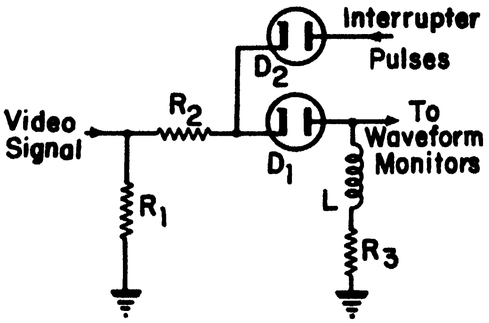
Fig. 18-47.—(A) Video signal obtained from detector circuit of Fig. 18-46; (B) signal obtained from circuit of Fig. 18-48.

Percentage Modulation Measurement

The video signal obtained from the detector circuit of Fig. 18-46 is shown in Fig. 18-47A. At any instant, the height ot the curve above the zero-voltage line corresponds to the instantaneous amplitude of the r-f signal. Hence, if the sync tips represent 100 per cent modulation, the modulation percentages corresponding to the blanking and white levels can be measured from the curve.

However, since the waveform monitors are a-c coupled devices, only the a-c component of the curve is shown on the screens; therefore, some method must be found for establishing the level of the zero-voltage line.

A commonly used method is to connect the contacts of a vibrator across the video line so that it is periodically short circuited; however, this system is not very satisfactory since the contact resistance of the vibrator is rather uncertain. Unless this contact resistance is small compared to 75 ohms, an effective short circuit is not obtained.

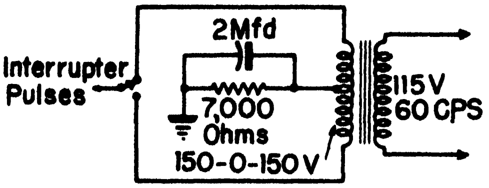


Courtesy Communications

Fig. 18-48.—Electronic method of signal development.

A more reliable electronic method is shown in Fig. 18-48 Since the video signal is negative with respect to ground, it is conducted by diode *D1* and fed to the monitor input, its amplitude being reduced in the ratio $\frac{R3}{R2+R3+R4}$ where *R4* is the diode a-c resistance.

The plate of diode *D2* is normally maintained at a potential more negative than the video signal so that it passes no current. At intervals, positive pulses are applied to the plate of *D2* and they are conducted by the tube and returned to ground via *R2* and *R1*. For the duration of a pulse, the cathode of *D1* becomes positive, *D1* ceases to pass current, and the voltage input to the monitors is zero. Hence, a series of interruptions occur in the video signal and we obtain the trace shown in Fig. 18-47B on the *frame* waveform monitor screen. The level of zero voltage is established by the bottom of the interrupting pulses. The a-c resistance of the diode *D1* will vary somewhat with voltage, so *R2* and *R3* are made as large as possible in order that the monitor input voltage shall be directly proportional to the video line voltage. As in the case of the monitor bridging circuit, a peaking coil *L* is connected in series with *R3* to correct the high-frequency response at the monitor input.



Courtesy Communications

Fig. 18-49.—Pulse-generating circuit showing change-over switch which shifts the phase of the pulses so that they can be prevented from coinciding with the vertical sync pulses.

It is convenient to make the pulse frequency equal to a multiple of the video *frame* frequency, since then the interrupting pulses remain stationary with respect to the sync pulses on the frame waveform monitor screen. This condition is met by the pulse-generating circuit of Fig. 18-49 as the frame frequency of the video signal is locked at half the a-c line frequency. The change-over switch in Fig. 18-49 shifts the phase of the pulses 180° so that they can be prevented from coinciding with the vertical sync pulses.

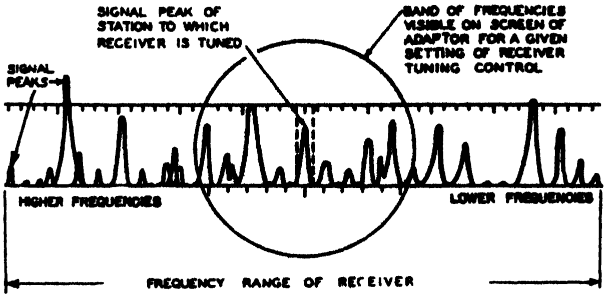
PANORAMIC TESTS[10]

Spectrum analysis is made possible in the range of radio frequencies by means of panoramic reception. A convenient piece of equipment for this purpose is the Panadaptor, Model PCA-2, to be used in conjunction with a standard superheterodyne receiver.

Panoramic Operation

Signals within a band extending up to 100 kc above and below the frequency to which the companion receiver is tuned appear as individual vertical pips, in order of frequency, at a definite location along the calibrated horizontal axis of the panoramic screen.

As the receiver is tuned, the pips move across the screen and "walk off" at one side while new pips enter on the opposite side. For any setting of the receiver tuning dial, the pip appearing at the *center* or *zero mark* of the screen represents the signal to which the receiver is tuned.

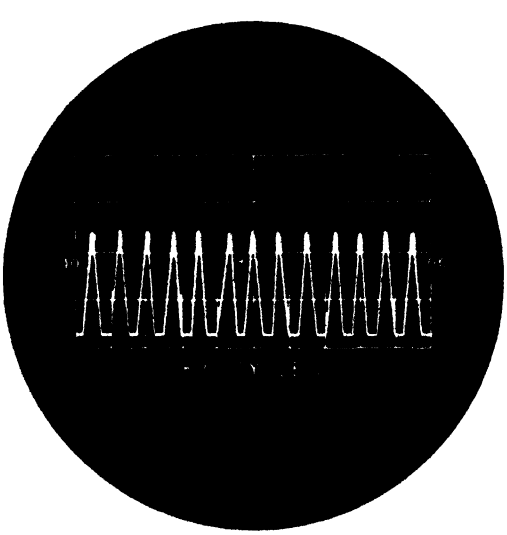

SIGNAL PEAK OF STATION TO WHICH RECEIVER IS TUNED
BAND OF FREQUENCIES VISIBLE ON SCREEN OF ADAPTOR FOR A GIVEN SETTING OF RECEIVER TUNING CONTROL
SIGNAL PEAKS
HIGHER FREQUENCIES
LOWER FREQUENCIES
FREQUENCY RANGE OF RECEIVER

Courtesy Panoramic Radio Corp.

Fig. 18-50.—Representation of a visible image obtained for a portion of the spectrum.

From the location of a pip and the receiver tuning dial setting, the frequency of the respective signal is determined. The height and behavior of a pip indicate relative signal strength and character, respectively.

When the Panadaptor is connected to some receivers with two or more stages of r.f., it may not be possible to observe up to 200 kc at one time. The visual bandwidth or sweepwidth can be reduced, at will, from 200 kc to 0 kc, continuously.



Courtesy Panoramic Radio Corp.

Fig. 18-51.—Picture of a signal shown in visible single-signal operation.

A representation of the visible image obtained for a portion of the r-f spectrum appears in Fig. 18-50. A typical picture of a signal shown in visible single-signal operation (overmodulated tone signal) appears in Fig. 18-51, while Fig. 18-52 is a typical panoramic picture of a 200-kc portion of a band; Fig. 18-53 illustrates the

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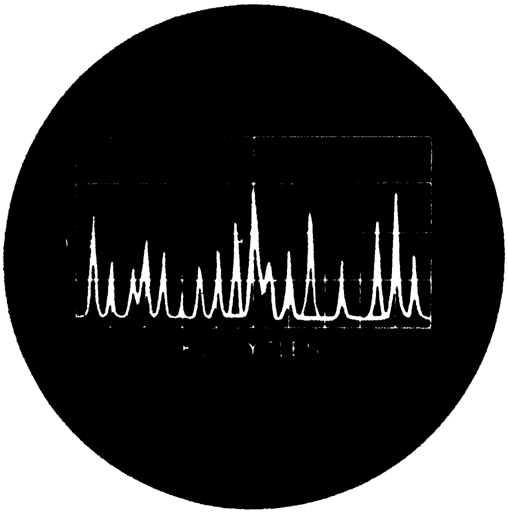
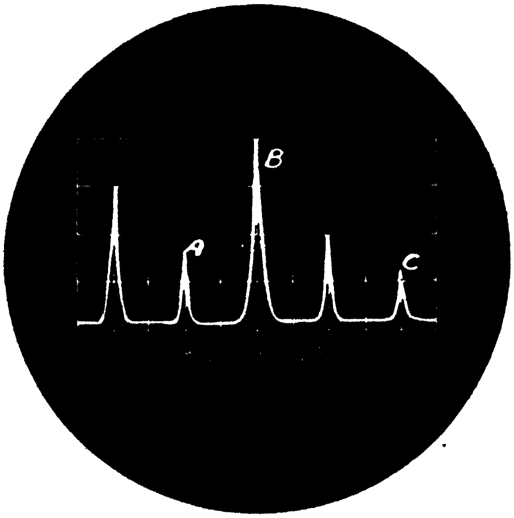
[10] The following material is an abridged excerpt from the instruction manual for the Panadaptor, Model PCA-2 of the Panoramic Radio Corp.



Courtesy Panoramic Radio Corp.

Fig. 18-52.—Typical panoramic picture of a 200-kc portion of a band.



Courtesy Panoramic Radio Corp.

Fig. 18-53.—Illustration of determination of signal frequency at maximum sweepwidth.

determination of signal frequency at maximum sweepwidth; at *A*, there is indicated a signal 40 kc above signal heard through receiver; at *B* is shown the signal heard through receiver; at *C* is a signal 80 kc below signal heard through receiver.

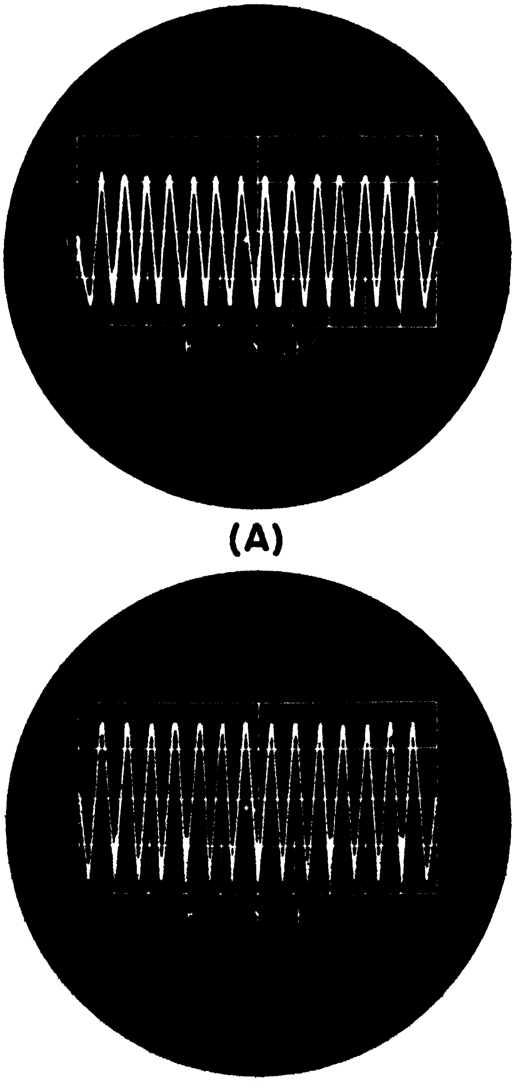
Signal Frequency Determination

As the receiver is tuned from a low to a high frequency, the signal deflections will move across the screen from left (+) to right (−). The reverse is true when tuning from a high to a low frequency. Those signals appearing on the (+) side of the zero mark are higher in frequency than the station heard through the receiver by the amount indicated by the screen calibrations. Each calibration mark, when the sweepwidth is maximum, is equal to 20 kc.

The above is true only when a local oscillator in the receiver tracks at a *higher* frequency than the incoming signal. The signs are reversed when the receiver local oscillator tracks at a lower frequency than the incoming signal.

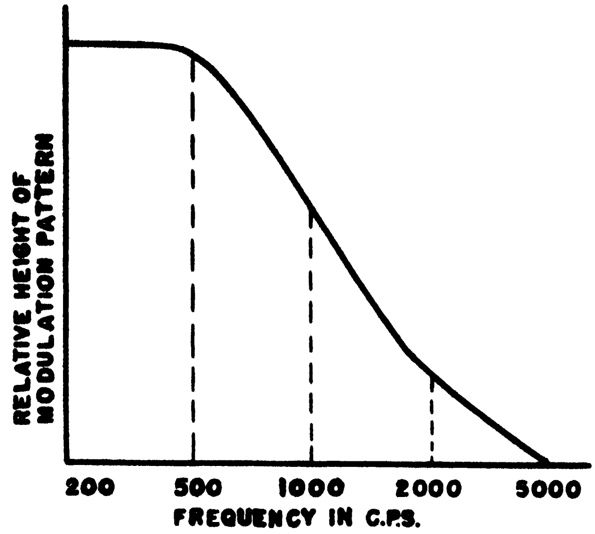
Measuring Percentage of Amplitude Modulation

When the Panadaptor is used at the transmitting end to examine the output of the transmitter, it may be of help to remove the antenna from the receiver or substitute a short length of wire for it, or even "short" the antenna and ground posts. The transmitter power may be reduced to prevent overloading of the receiver.



(B) *Courtesy Panoramic Radio Corp.*

Fig. 18-54.—Illustration of cathode-ray-tube screen calibration for modulation measurements.



Courtesy Panoramic Radio Corp.

Fig. 18-55.—Variation of waveform height (or calibration) against modulating frequency at 100 per cent modulation.

Before measurement can be made, it is necessary to calibrate the panoramic screen, in terms of percentage of modulation, against a source of amplitude-modulated r.f. in which the modulation percentage can be determined. A good signal generator with a reliable modulation meter or any station either local or distant with facilities for measuring percentage of modulation of its carrier will do.

The Panadaptor is first set for "Visible Panoramic" and the receiver is tuned to a signal, modulated by a frequency under 500 cps, from either of the above sources, and then operation is switched to "Visible Single Signal." The "Center Frequency" control or the receiving tuning dial is trimmed for maximum deflection. The baseline is elevated by the "Gain" control so that it coincides with the center horizontal line on the screen. It may be necessary to remove modulation to facilitate this adjustment.

Beginning with zero modulation, the modulation percentage is increased in steps of 20 per cent or any other desired percentage. The height of the modulating waveform on the tube screen is noted for each step and the screen is calibrated. The upper half of the modulation is limited by pulse agc and will appear somewhat shorter than the lower half. This is illustrated in Fig. 18-54. It is important that the modulation be less than about 500 cps for accurate modulation depth readings. The graph in Fig. 18-55 shows a typical variation of waveform height (or calibration) against modulation frequency at 100 per cent modulation.

Overmodulation is indicated by flattening or clipping of the lower peaks of the modulating waveform along the baseline.

Checking and Adjusting Frequency Deviation of F-M Signals

It is possible to check and monitor the frequency deviation of an f-m signal instantaneously at the receiving or transmitting end, by observing it on the panoramic screen.

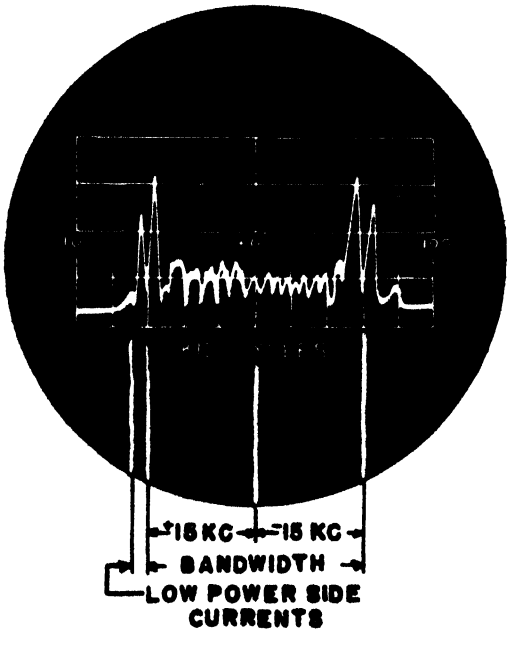
For voice or music f-m signals, the extent of frequency deviation (and, therefore, audio level) varies continuously; whereas, for a tone f-m signal of uniform strength, the deviation is constant. For either case, the extent of deviation is instantly shown by the limits of the horizontal spread of pips across the screen.

The bandwidth of the signal is equal to the sum of the frequency deviation to the right and left of the center zero mark on the screen.

The signal energy distribution of the signal through the frequency spectrum is shown by the envelope of the pips.

To check frequency deviations within narrow limits such as ±15 kc, the "Sweepwidth" control is adjusted to give a maximum visual display of 50 kc. At this setting, each calibration mark on the screen equals 5 kc.

It is important that the gain be sufficiently high so that the sideband "pips" at the ends of the spread are



Courtesy Panoramic Radio Corp.

Fig. 18-56.—Instantaneous deviation of ± 15 kc from the center frequency.

shown. An instantaneous frequency deviation of ±15 kc from the center frequency is shown in Fig. 18-56.

It will be noted that pips which are approximately 4 kc apart tend to merge below their waists. This is normal and the energy distribution is still shown by the envelope of the pattern.

If a broadcast f-m station is monitored for a 150-kc maximum bandwidth, the "Sweepwidth" control should be at maximum. The appearance of modulation pips outside the 75-kc limits on either side of the zero mark on the screen indicates overmodulation.

To adjust the upper limit of modulation of an f-m transmitter, it is advisable to use a constant tone modulating voltage from an a-f tone oscillator. A tone frequency of 15 kc is suggested for broadcast f-m stations.

It will be noted that when a 15-kc audio is used and the audio gain control of the modulator is advanced up from zero, the height of the carrier deflection, on the screen will gradually fall and pips will appear and rise 15 kc on either side of center.

As the modulation is increased, two new pips will appear 30 kc on either side of zero. As they rise the carrier will fall to zero level (disappear) when the frequency deviation is 2.4 times greater than the audio modulating frequency.

Further increases of modulation result in the appearance of new pips 45 kc, 60 kc, 75 kc, etc., on either side of the center; and in passing through given deviation ratios, the carrier and pips successively fall to zero and then rise, as was discussed earlier in the chapter.

A practical and accurate means of obtaining a particular frequency deviation is to set the a-f tone oscillator for a frequency equal to the particular frequency deviation desired divided by a deviation ratio for zero carrier level and advance the audio gain control until the carrier deflection goes to zero the proper number of times. *Example*: For a deviation of 60 kc, the audio oscillator is set at $60/2.4 = 25$ kc and the audio gain control (modulator) is adjusted until the carrier deflection dips to zero once.

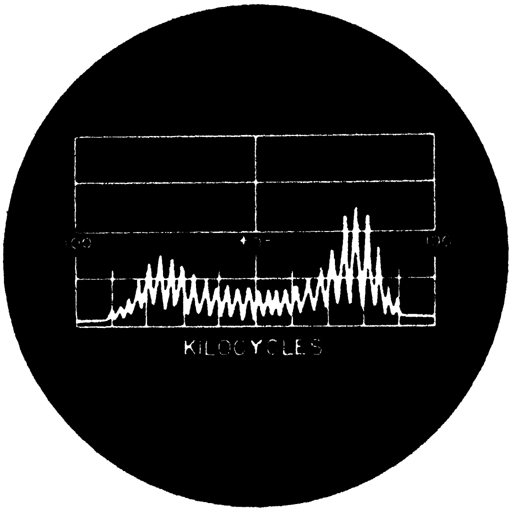
For $60/5.52 = 10.87$ kc, the gain control is advanced so that the carrier dips to zero twice.

Identifying A.M. on F-M Signals

Normally, signals which are frequency-modulated by a sine wave consist of vertical deflections equally disposed on both sides of the carrier. The pattern of the envelope of the deflections is, therefore, symmetrical about the carrier.

The presence of a.m. on an f-m carrier is indicated by a twist in the envelope pattern. The vertical dimensions of the pattern will appear larger on one side of the carrier-pip than on the other side, as shown in Fig. 18-57

In examining a received or transmitted signal, the Panadaptor is operated on "Visible Panoramic." It is preferable that the station be modulated by a steady sine wave so that a stationary pattern of deflections is obtained.



Courtesy Panoramic Radio Corp.

Fig. 18-57.—Presence of a.m. on an f-m carrier. The vertical dimensions of the carrier appear larger on one side of the carrier pip than on the other.

Detecting Residual Hum on a Carrier

The carrier to be examined is tuned to the center of the tube screen and the Panadaptor is adjusted for "Visible Single Signal." If the elevated baseline is curved or hooked, it indicates hum on the carrier. Another carrier may be tuned in and its baseline can be compared with the former to determine the relative hum content.

If it is f-m hum, the phase of the hum on the baseline is reversed as the "Center Frequency" control is passed through its zero position or if the receiver is tuned through the station slowly. But if it is a-m hum, this phase reversal does not occur.

CHAPTER 19

ELECTRICAL MEASUREMENTS, SCIENTIFIC AND ENGINEERING APPLICATIONS

This chapter covers the standard applications of the cathode-ray oscilloscope in electrical measurements, such as those used in radio work, and the scientific and engineering applications which, by comparison, are specialized and more involved.

Electrical Measurements

Electrical measurements fall into a number of classifications. Perhaps the most fundamental measurement is that of current. In measuring voltage we are really measuring a current, since most indicating instruments operate from the magnetic effect of a current; the resistance of the measuring circuit is such that the current and the voltage to be measured are in a particular ratio.

However, it has become customary to speak in general terms of voltage, current, impedance, and power measurements, as these are the quantities ordinarily required to be measured in radio and electronic work. The greatest utility of the oscilloscope is in making alternating voltage measurements. The measurements may be in absolute or general terms. For example, knowing the deflection sensitivity in volts per inch, the equivalent voltage for fractions of an inch, or multiples of such fractions, can be found.

In the usual applications, the shape and relative height of the trace on the screen of the cathode-ray tube are the only important factors. However, the cathode-ray oscilloscope can be used for the measurement of electrical quantities and its uses in such applications will be described in the following discussions.

Since there is no fixed calibration, the cathode-ray oscilloscope is not particularly suited to absolute scale measurements of voltage, either a-c or d-c, but rather to portrayals of the *waveforms* of voltages. The shape of the waveform curve is indicative of the condition or quality of the equipment under test. The oscilloscope can be used, with fair accuracy, as a voltmeter, provided a calibration is made. It has the great advantage of good transient response and low inherent inertia.

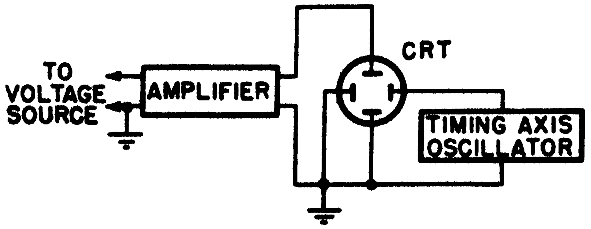
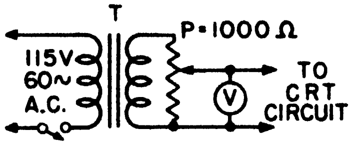


Fig. 19-1.—Block diagram of fundamental a-c voltmeter circuit.

A-C Voltmeter

A block diagram is shown in Fig. 19-1. This is the usual oscilloscope arrangement. The timing axis oscillator is usually adjusted to the frequency of the incoming signal for a single trace. The timing axis oscillator may sometimes be adjusted for ½, ⅓, or ¼ of the frequency of the voltage. Obtaining a single trace is desirable in the interests of simplicity.

Fig. 19-2. — Calibrating circuit for cathode-ray oscilloscope using a-c voltmeter.



If the oscilloscope input is connected to a 1-volt, 60-cps calibrating source, such as that shown in Fig. 19-2, it will be accurate as an a-c voltmeter over a range corresponding to the flatness of response of the vertical amplifier. For example, if a 1-volt, 60-cps signal produces a 0.1-inch deflection, and the oscilloscope amplifier is rated as flat to 100,000 cps, the 0.1-inch deflection will hold true for a voltage of any frequency up to 100,000 cps.

Some oscilloscopes, of special design, are virtually flat, linear, in response from d.c. to as high as 10 Mc. This feature is a great advantage in a voltmeter. Further, the input impedance of the oscilloscope is usually quite high, seldom less than a megohm. This means that the circuit loading of the oscilloscope type of voltmeter will usually be negligible in all but very high impedance circuits. However, even here, the impedance loading may be minimized using low-capacitance shielded test cable and a circuit-isolating type of probe. An arrangement of this sort is shown in Fig. 19-3.

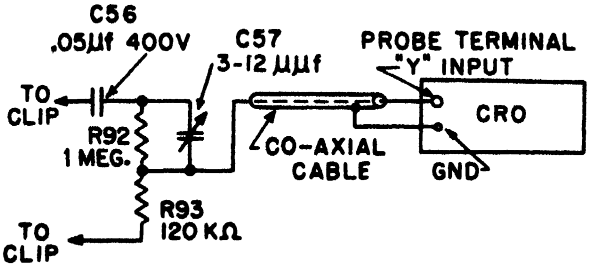


Fig. 19-3.—Probe input circuit for cathode-ray oscilloscope.

As the frequency is raised to a point beyond the range of the amplifier, the oscilloscope voltmeter becomes ineffective. The frequency range may be still further advanced, with a corresponding loss of sensitivity, by removal of the amplification system interposed between the vertical-deflection plates and the test circuit. The timing axis oscillator circuit usually presents no particular problems, since it is not a difficult matter to design stable oscillator systems capable of producing sweep signals up to the highest frequencies necessary. Most oscilloscopes, particularly the better types, have some provision for direct connection of the test circuit to the vertical plates. External horizontal plate terminals are also usually afforded.

In connection with Fig. 19-2 and calibration of the circuit, while there may be quite a difference between the voltage looking into the probe and that at the output of the probe (oscilloscope input), the circuit will be useful for comparison measurements and approximate high-frequency voltage measurements. In the Du Mont 241 oscilloscope, for example, the *Y*-axis input impedance is 2 megohms in parallel with 40 *μμf*, without the probe. With the probe, it is 1 megohm and 10 *μμf*. Using direct connections to the vertical plates (unbalanced), the 241 oscilloscope input impedance would be 5 megohms in parallel with 25 *μμf*. The balanced input impedance would be about the same value.

Mention has been made, previously, of a deflection on the *Y* axis (vertical) of 0.1 inch for 1 volt. Up to the input overloading level of the oscilloscope *Y*-axis amplifier, the deflection will be linear and proportional to voltage. For example, for the case mentioned, if 1 volt = 0.1 inch, then 0.2 inch = 2 volts, 0.3 inch = 3 volts, etc. Some manufacturers may specify the deflection sensitivity in millimeters per volt d.c. or rms rather than in inches per volt.

The range of the oscilloscope voltmeter may be extended by using a suitable series multiplier resistor, just as it is in the case of an ordinary voltmeter. However, because the input impedance of the oscilloscope is so high, and the current flowing in it so small, the voltage rating and the physical size of the series resistor can be kept down. This is a distinct advantage. Further, the use of the multiplier increases the oscilloscope effective input impedance and reduces the loading of the test circuit. A fundamental requirement of any measuring instrument, and the oscilloscope is no exception, is that it load the circuit under test as little as possible.

Most oscilloscope *Y*-input circuits have a series capacitor which blocks d.c. and passes a.c. Therefore, the input circuit of Fig. 19-1 may usually be directly connected to any d-c source having a superimposed a.c., without damage to the oscilloscope, provided the d-c potential is no higher than 300 or 400 volts. The maximum safe input potential is often, but not always, stated by the oscilloscope manufacturer, in the instruction manual. In the plate circuit of an audio power output tube in a radio receiver, the d-c potential might be 250 volts. This is illustrated in Fig. 19-4. *L* is the primary inductance of the output transformer primary, *C* is the usual low-impedance capacitor of the power supply, and *VT* is the output tube. It makes no difference whether the output stage is single-ended or not;

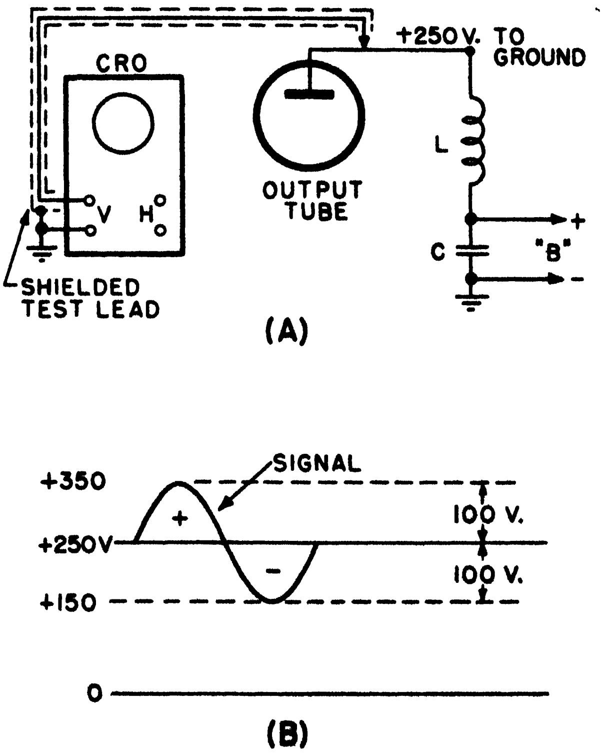


Fig. 19-4.—Connection of the oscilloscope in checking for hum. A 600-volt 0.1-*μf* capacitor may be used in series with the "hot" side of the circuit if the oscilloscope does not have a built-in blocking capacitor.

either output tube circuit of a push-pull or parallel push-pull system could be checked. In the drawing, a sine wave having a 100-volt peak value is shown. Then, on positive peaks the maximum input to the oscilloscope is 350 volts and on negative peaks the minimum is 150. The signal voltage to ground, indicated by the instrument, is 100 volts peak.

Most oscilloscope inputs have positive d-c polarity; that is, there may be a momentary upward kick of the horizontal timing axis line as the oscilloscope is connected to the circuit under test, when the polarity of the d.c. is such that the ungrounded terminal of the oscilloscope input is + with respect to the grounded terminal. Similarly, with an applied a-c voltage, the deflection will be upward when the ungrounded side is going through a positive peak. Some have adjustable polarity or negative polarity. This effect is normal. By setting the vertical gain control at low gain initially, and then bringing it up to the required setting, possible overloading of the oscilloscope is avoided.

If there is any doubt about the input circuit of the oscilloscope, a 0.5 μf 600-volt series capacitor may be used externally, located close to the vertical input terminal, to block direct current and pass the alternating current. The oscilloscope, as an a-c voltmeter, may be connected in the receiver or electronic equipment circuit in place of the usual copper-oxide rectifier — d-c voltmeter combination ordinarily used for a-c voltmeter applications. The oscilloscope has the advantage of showing the peak value and waveshape of the voltage. Incorrect operation, or poor design, are often indicated by distorted waveshapes. The ordinary voltmeter gives no information on waveform and may read inaccurately in the presence of harmonics. Also, its frequency range is limited and temperature changes may adversely affect its accuracy. The oscilloscope, on the other hand, is relatively free of such defects.

With respect to Fig. 19-4, the oscilloscope input may be connected across the output filter capacitor *C* to check the relative efficiency of the capacitor and filter. A serviceman working in a noisy radio shop may find it very difficult to check the relative hum level of a receiver. The local noise of the shop and the fact that the loudspeaker is not available or not mounted in its cabinet (baffle) will result in test conditions that do not parallel those found in the home. A quantitative measurement of hum voltage, therefore, is valuable. If the power supply uses a 60-cps line source, the ripple frequency will be 60cps, and the timing axis oscollator may conveniently be adjusted to 60 cps. Even if the set uses a full-wave rectifier, it is better to use the 60-cps sweep, to show both halves of the ripple for comparison.

In checking power-supply filters, the usual practice of servicemen is to shunt a test capacitor across one suspected of being faulty and listen to the effect upon receiver or amplifier performance. The great advantage of this method is speed. While it may be somewhat slower, a better technique is to unsolder one lead of the old capacitor, connect a new unit into the circuit, and make a comparison of the a-c voltage output magnitude and waveform, using the oscilloscope. Hum will be treated in detail later on, but it may be noted here that in its role as an a-c voltmeter, a basic function, the oscilloscope can be quite useful and a distinct timesaver in circuit analysis.

The value of the voltages will vary with different receivers and amplifiers. The relative deflections will vary with the type of oscilloscope and its sensitivity. Therefore, it is a matter of the individual user becoming thoroughly versed in the use of his particular instrument and acquiring experience on many sets and amplifiers, to derive the greatest possible benefit from the oscilloscope as a servicing or engineering aid.

D-C Voltmeter

The oscilloscope can be used as a d-c voltmeter provided a direct connection is made to the vertical plates or a d-c amplifier is used between the vertical plates and the circuit under test. As d-c amplifiers are, for various technical reasons, rarely used, the usual procedure is to employ the direct connections.

It is possible to ground the two horizontal plates and one vertical plate, then apply the d-c potential across the vertical plate circuit and thus obtain a movement of the spot position corresponding in magnitude to the relative value of the d-c voltage under test. This method is illustrated in Fig. 19-5. The movement of the spot is upward on the vertical or *Y* axis and dependent on the size of the cathode-ray tube, the general tube and cathode-ray-tube circuit design, and the value of the voltage. The cathode-ray tube can be calibrated with a d-c voltage and potentiometer as shown in Fig. 19-6. The voltage source may be a storage battery, a

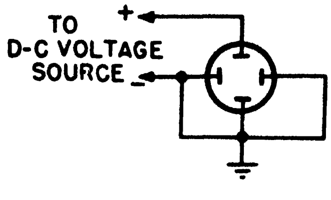


Fig. 19-5.—D-c cathode-ray oscilloscope voltmeter connections.

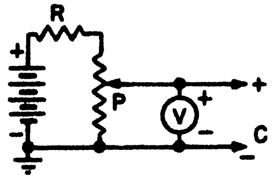


Fig. 19-6.—D-c calibrating circuit for cathode-ray oscilloscope.

group of "B" batteries, or a regulated power-supply system, depending on the voltage range desired. The potentiometer *P* must be capable of handling the power in the circuit. A practical potentiometer or voltage divider may take the form of a wire-wound resistor equipped with an adjustable slider. For a 300-volt source, a 25,000- to 50,000-ohm, 20-watt unit would be considered practical. Resistor *R* should be of such value that reasonable protection is afforded the cathode-ray tube, with the slider on *P* adjusted to maximum voltage; the safe input voltage rating of the cathode-ray tube must not be exceeded. For some applications, *R* will not be needed at all. For others, its inclusion is necessary. *V* is a standard d-c voltmeter which may take the form of a laboratory instrument.

The use of a stationary spot, however, may possibly result in burning of the screen. If a sawtooth sweep voltage is applied to the *X* plates in the conventional manner, a straight line will be obtained across the face of the tube. This horizontal bar can be moved up or down just as conveniently as a spot and with far less danger to the screen. The sweep frequency is not critical and may be set conveniently at any arbitrary value, perhaps 400 or 1,000 cps. The d-c voltage is then applied to the vertical plates, the sweep to the horizontal, and the vertical movement of the horizontal line or bar is proportional to the applied d-c potential.

Strays

It is essential that the lead used to connect the oscilloscope input circuit to the circuit under test be properly shielded to avoid stray pickup. Using low-impedance circuits, in general, tends to minimize pickup of stray hum and noise, but the nature of the usual oscilloscope setup is such that a high-impedance input circuit is required.

Current

The measurement of current is not ordinarily the function of a cathode-ray tube, which, basically, is a voltage-actuated device. Current can be measured indirectly by measuring the voltage across a resistor in series with the circuit under test. The current is then equal to *E* divided by *R*, using the familiar Ohm's law equation. The oscilloscope can be calibrated in a number of ways, depending upon the type of measurement, a.c. or d.c. The advantage of the oscilloscope is its wide frequency range and its ability to show the ***waveform*** of a current rather than its absolute ***magnitude***. Fig. 19-7 shows an obvious arrangement for current calibration. *R* is low in comparison with the test circuit resistance.

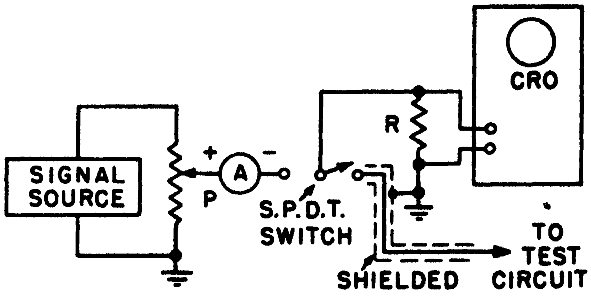


Fig. 19-7.—Circuit for calibration of cathode-ray oscilloscope as an ammeter.

Alternating Current

A circuit for a-c measurements and calibration is shown in Fig. 19-8. The signal source may be a variable frequency audio generator capable of being set to any desired frequency, or the secondary of a small transformer supplying a few volts of 60-cps frequency. Output potentiometer *P* may be the audio output control of the generator or a potentiometer shunted across the transformer secondary. *E* is a standard a-c voltmeter, either of the copper-oxide rectifier type or a vacuum-tube voltmeter. *S*, the single-pole double-throw switch, is thrown to the left, for calibration purposes, and to the right for current measurements. *R* is a resistor of any convenient size, low enough in comparison with

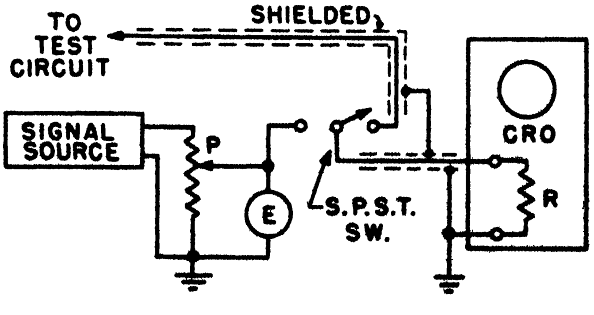


Fig. 19-8.—Circuit for a-c measurements and calibration.

the load resistance of the device being tested to have negligible insertion effect. *R* may be an adjustable wire-wound resistor for low frequencies, but preferably should be a noninductive type.

A somewhat similar arrangement, except that an ammeter, rather than a voltmeter, is used, is shown in Fig. 19-8. However, few ordinary commercial generators are able to deliver very much current to a low-impedance load and this circuit, therefore, is ordinarily less satisfactory than the one that preceded it in this discussion.

A-C Power

The apparent power in an a-c circuit is the product of voltage and current. The true power is less than the apparent power for all but a purely resistive load. The true power is $EI \cos \theta$ where θ is the phase angle. Therefore, if we know E and I and find $\cos \theta$, the power factor, we can determine the true a-c power by using the formula. The angle θ can be found from a slide rule or by consulting a trigonometric table, knowing that $\sin \theta = A/B$, where A and B are as defined in Fig. 19-9. The value of $\cos \theta$ can then be found at once; it is the power factor. The sum of the resistances *R1* and *R2* is made equal to the plate resistance of the tube used with the loudspeaker or load impedance Z, and the voltage E is made equal to the equivalent generator voltage of the tube, μe_g, where μ is the amplification factor of the tube and e_g is the signal voltage acting in the grid circuit.

P is a potentiometer of convenient size for the source voltage available. *T* is the usual output transformer with the high-impedance side connected to the load and the voice coil or low-impedance side connected to the source of driving voltage. *E* is a standard a-c voltmeter.

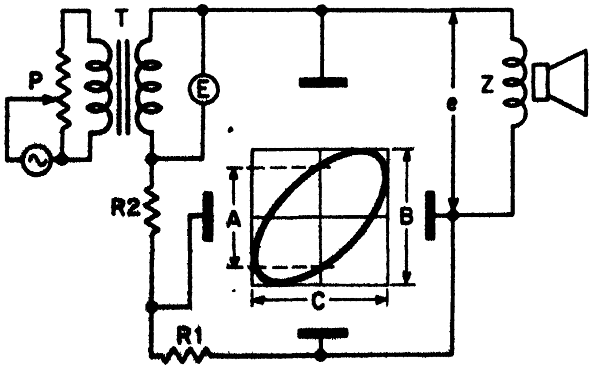


Fig. 19-9.—Circuit for measurement of phase angle from which power factor, apparent power, and real power may be calculated.

This method of measurement utilizes the relationship between the vertical- and horizontal-deflection sensitivity of the cathode-ray tube and makes calculations of the impedance and power factor independent of acceleration voltages used on the anodes of the cathode-ray tube. K is the ratio of the vertical-deflection sensitivity in direction B, Fig. 19-9, to the horizontal-deflection sensitivity in direction C. The phase angle, $\sin \theta = A/B$. The frequency and magnitude of E is varied as desired, and the resulting impedance and power factor, $\cos \theta$, are determined from the formula given above, and $Z = \frac{KBR_1}{C}$.

D-C Power

In a d-c circuit, power is simply EI. The power can be found, therefore, by measuring these two quantities and multiplying them together. Two separate oscilloscopes and circuits could be used, or a single oscilloscope and switching arrangement. Fig. 19-10 shows a system of this kind. *R* is a resistor of some convenient value. It is assumed that the oscilloscope has been calibrated for E and I, using the methods previously described; then, $P = EI$, a simple product.

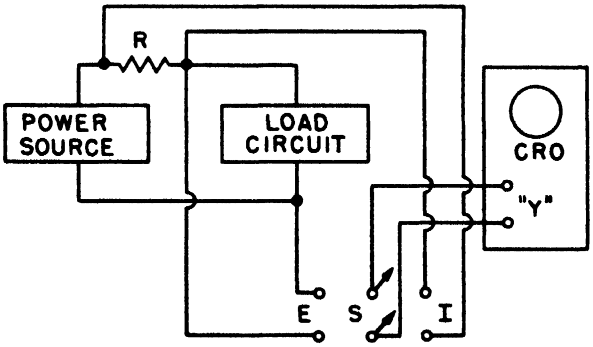


Fig. 19-10.—Measurement of d-c power is possible with this arrangement.

Magnetic Circuit Tests

Measurement of hysteresis, a magnetic circuit effect, is feasible using the cathode-ray oscilloscope. The fundamental circuit, for test purposes, is illustrated in Fig. 19-11. While the radio technician who services receivers would not be called upon to make hysteresis measurements, the electrical fundamentals involved are important to him and a knowledge of them is essential to a complete understanding of his work. To the

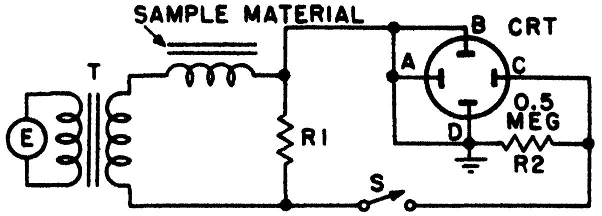


Fig. 19-11.—Basic circuit for hysteresis measurements.

scientist or engineer, magnetic circuit measurements are of great practical importance and general utility in conducting investigations into the behavior of circuit components.

The general theory of magnetic circuits is given in standard texts on the subject. Our purpose here is to present an application of the cathode-ray tube at work. The circuit is so arranged, in Fig. 19-11, that the flux density establishes the vertical deflection, as with ordinary electromagnetic deflection oscilloscopes. The flux is produced by the flow of alternating current of the required frequency and sufficient intensity to provide a suitable field strong enough to deflect the beam. Consequently, the vertical deflection indicates *B*, the flux density. At the same time, the circuit provides for a deflection at right angles to that of the flux density, proportional to the magnetizing current. This deflection is developed through application of the voltage across *R1* to plate *C* of the cathode-ray tube and ground. The magnetizing current and voltage across this resistor are in phase. The sample core material is placed within the solenoid and held near the neck of the cathode-ray tube, with the face of the solenoid parallel to the plane of the horizontal-deflection plates. Since the deflection of the beam is at right angles to the magnetic field, positioning of the testing solenoid will produce the required vertical deflection. The switch marked *S* is normally closed. The switch is included in the circuit so that, if desired, the two axes may be produced on the cathode-ray-tube screen for photographic purposes. When switch *S* is open, the vertical deflection alone is on the screen. To produce the horizontal deflection only, without the vertical deflection, switch *S* is closed and the solenoid and core sample are bodily removed from the neck of the tube.

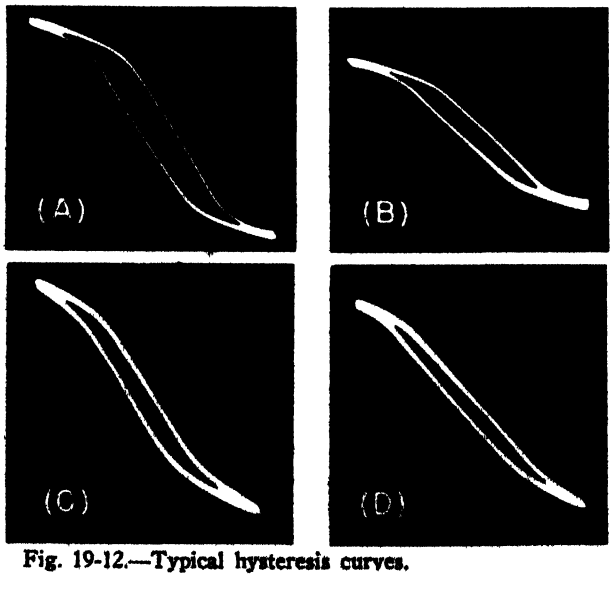


Fig. 19-12.—Typical hysteresis curves.

Illustrative *B-H* curves are shown in Fig. 19-12A, B, C, and D. These curves are those obtained on the cathode-ray-tube screen without vertical and horizontal axis lines. They were made at a frequency of 100 cps. The value of the resistor *R1* is dependent upon the signal voltage available. This is due to the fact that, with limited input signal voltage, saturation of the core, indicated by departure from a straight line, is controlled by the current-limiting action of this resistor. To show the saturation effect, very low values for *R*, as low as 10 or 12 ohms, were used. However, with higher values of input voltage, so that saturation can be established by varying the input signal voltage, the value of this resistor may conveniently be 100 ohms or larger.

The ideal **B-H** curve is a single straight line. Saturation of the core, with its detrimental effects, is indicated by the flattening of the ends of the loop, as indicated in Fig. 19-12A, B, C, and D. The most pronounced case of saturation is that in Fig. 19-12A. Cases of reduced saturation are indicated in Fig. 19-13A and B. These patterns also indicate the core loss, essentially due to hysteresis. (This means the loss introduced in the form of heat, because of the molecular change which takes place during the process of magnetization.) The extent of this loss is indicated by the area of the loop. The greater the area of the

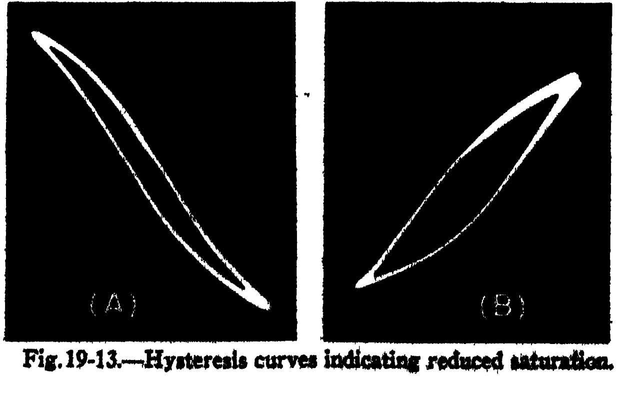
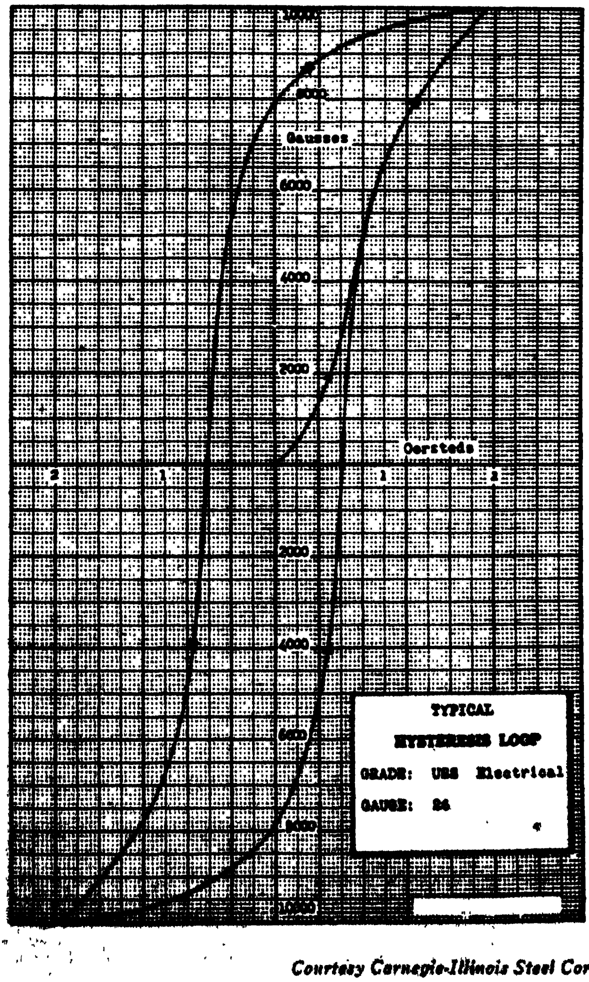


Fig. 19-13.—Hysteresis curves indicating reduced saturation.

hysteresis pattern, the greater the hysteresis loss indicated in each of the patterns, but Fig. 19-13B shows the greatest amount of core loss due to hysteresis, yet minimum amount of core saturation.

When making such magnetic measurements, the sweep circuit is not used. The cathode-ray oscilloscope is adjusted so that independent voltages may be applied to the vertical and horizontal plates through the amplifiers feeding these deflection plates.

In the event that the cathode-ray oscilloscope is not equipped with amplifiers for the two sets of deflection plates, or for either set of deflection plates, it will be necessary to apply sufficient signal voltage to the sample coil and rheostat circuits so that suitable deflections are obtained, despite the absence of the amplifiers. The test frequency is not limited to 100 cps. It can be whatever frequency is desired, in accordance with the requirements of the test. Transformer steel manufacturers publish **B-H** curves, such as that shown in Fig. 19-14, which can be referred to when making tests.



Courtesy Carnegie-Illinois Steel Corp.

Fig. 19-14.—Typical transformer steel B-H curve.

Tube Characteristics

Tube characteristic curves can be shown advantageously, for educational or engineering purposes, by means of the cathode-ray oscilloscope. However, the basis of the curve technique must be understood before new curves can be developed or standard curves utilized. Electrical fundamentals continue to apply in this work, by extension of basic theory to cover the specialized application. The various classes of tubes may roughly be divided into high-vacuum and gaseous types. Each of these may be divided into generic types — diode, triode, tetrode, pentode, etc., according to the number of elements contained in the tube and its general construction.

Diodes

The diode may be used as a d-c limiter, audio rectifier, power rectifier, or r-f rectifier. According to its application, an r-f rectifier may be termed demodulator, mixer, or detector, but here we are concerned only with some very fundamental considerations relating to the applied voltage and the current that flows as a result of it. The shape of the curve will be of interest to us in picturing the action of the device.

The basic property of value in the diode is its ability to restrict current flow to one direction; thus, it may serve as a gate for electronic current, or as an electronic switch. In Fig. 19-15A, the current is limited only by the diode effective resistance and is from cathode to plate. In Fig. 19-15B, a resistor is in series

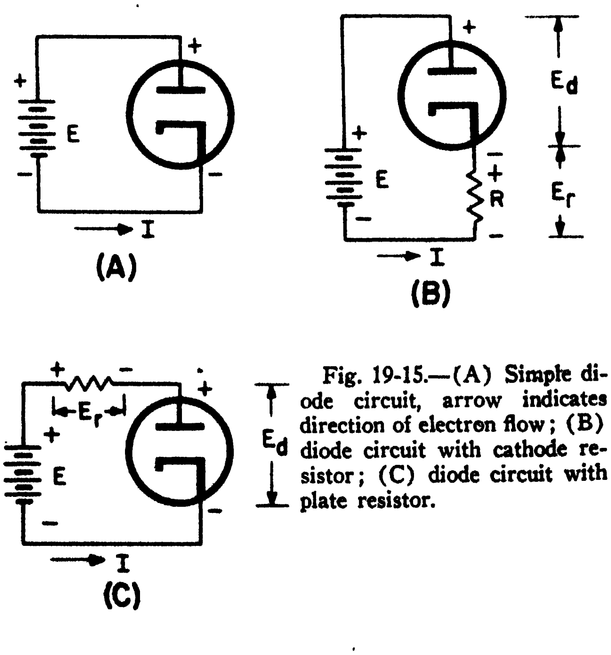


Fig. 19-15.—(A) Simple diode circuit, arrow indicates direction of electron flow; (B) diode circuit with cathode resistor; (C) diode circuit with plate resistor.

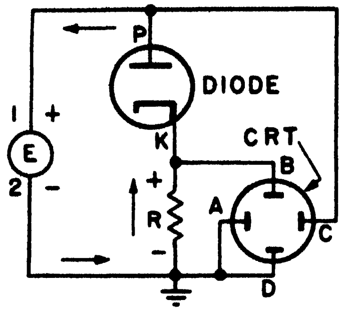


Fig. 19-16.—Connection of diode circuit to the cathode-ray tube for test purposes.

with the cathode, and the polarity of the voltage across it is indicated. The sum of E_d and E_r is equal to the source voltage E. In Fig. 19-15C, the same principle holds true, but the resistor is in the plate circuit, with the polarity shown.

If E is replaced with an alternating voltage, the diode will serve as a rectifier, conducting during the half of the a-c cycle when the plate is positive with respect to the cathode and not otherwise. Suppose, now, that we use such a source and connect it to the diode circuit and oscilloscope setup as shown in Fig. 19-16. We shall obtain a trace on the cathode-ray-tube screen indicative of the circuit behavior.

We may assume, for the purpose of circuit analysis, that point *1* is positive during one half of the a-c cycle and that a current flows in the direction indicated by the arrows. Then, plate *B* of the cathode-ray tube becomes positive with respect to *D* and a force acts on the electron beam toward *B;* simultaneously, due to the fact that *C* is also at a positive potential to ground, and with respect to *A,* there is a force acting to the right, and the resultant movement of the beam is the vector sum of these two forces. A typical curve obtained with half of a 5Y3 rectifier diode is shown in the Fig. 19-17. The tilt of the curve is due to the electromotive forces described.

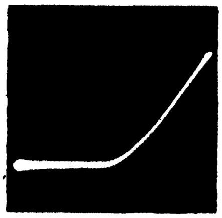


Fig. 19-17.—Typical characteristic curve of one-half of 5Y3 rectifier diode.

But what of the action during the opposite half of the cycle when the polarity reverses? Then, the polarity signs are reversed and the current flow is in a direction exactly opposite to that shown by the arrows, if there is any leakage through the diode. Assuming that this current is negligible, or does not flow at all, we have observed the functioning of a switch — the diode — which is now in the open position. The potential acting in the circuit makes *C* negative and both *A* and *D* positive; but since there is no current through *R,* and therefore no voltage drop, *B* is likewise at the same potential as *D*. The deflection force is that acting between *A* and *C*. Therefore, essentially the beam tends to move to the left, in a horizontal plane; this effect is shown in the flat portion of the curve in Fig. 19-17. The same fundamental principles would hold for any diode, at any frequency and any voltage. However, at ultra-high frequencies, for example, other complications such as transit time of the electrons would affect the behavior of the circuit in a manner that can not be explained simply. At the usual audio, power, and radio frequencies, the analysis is substantially true.

We have seen previously the behavior of d-c and a-c diode circuits. The functioning of a diode under the conditions of d.c. plus superimposed a.c. is an interesting study. Fig. 19-18A shows the basic test setup used. In an experiment conducted in the John F. Rider Laboratories, using one diode unit of a standard 5Y3, the curve shown in Fig. 19-19 was obtained. The diode plate is held at a positive potential at all times by the d-c source, whose voltage is greater than the peak value of the superimposed a-c potential. This is the usual case for many practical circuits found in radio work; therefore, the analysis should be particularly

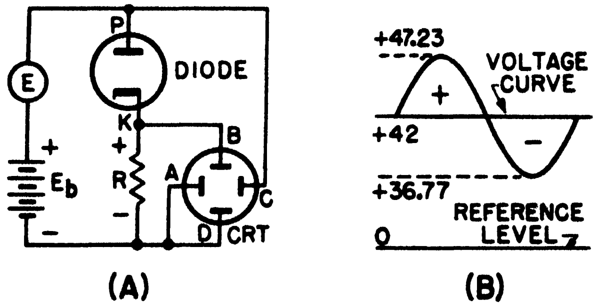


Fig. 19-18.—(A) Functioning of a diode under conditions of d.c. plus superimposed a.c. is studied with this circuit. The voltage relationships are shown in (B).

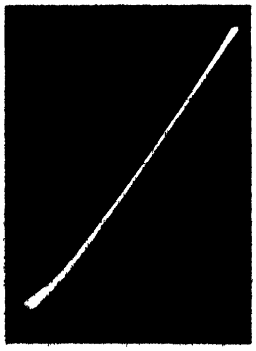


Fig. 19-19.—Characteristic curve of 5Y3 diode under test setup of Fig. 19-18A.

useful to practical men and engineers engaged in development work on such circuits.

The voltage relationships are shown in Fig. 19-18B. A diagonal line is obtained on the oscilloscope (Fig. 19-19). Its tilt and general characteristics will be dependent on the applied potentials and the value of R, as well as the diode characteristics. The curve shown in Fig. 19-20 was obtained with 51,000 ohms.

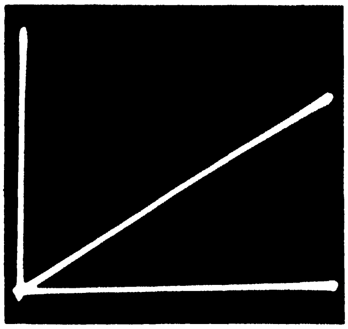


Fig. 19-20. — Diode (5Y3) characteristic curve obtained with series resistor of 50,000 ohms.

Triodes

The triode is simply a diode with a grid added. The grid affords great control over the plate current, with a small controlling potential. Fig. 19-21 shows the relationship that exists between the various factors of tube plate potential, grid bias, and plate current. For some particular value of plate potential X on the X axis, a vertical line perpendicular to X may be erected, so that it intersects the curve, which is determined by the particular bias potential used. Then, by drawing a horizontal line from the point of intersection to the Y axis, and finding point Y, the current in the plate circuit is found. When several different fixed values of grid bias are used, while the plate-voltage plate-current relationship is determined for each bias, a family of curves results. Such a family is shown in the photograph, Fig. 19-22. The circuit setup, used to obtain it, is illustrated in Fig. 19-23.

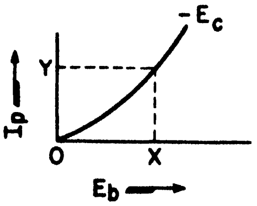


Fig. 19-21. — Graphical relationship of plate potential, bias, and plate current of vacuum tube.

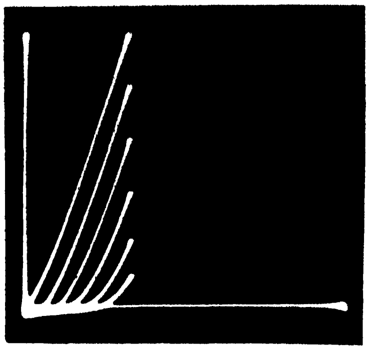


Fig. 19-22. — Family of triode e_g-i_p characteristic curves.

In the experimental setup, an alternating potential of 125 volts peak (88 volts rms) was used in series with a d-c potential of 125 volts, giving an applied voltage which alternated between zero and 250 volts. The bias values were adjusted by means of potentiometer P in the grid circuit of the triode (a 6J5) from zero to a maximum of -10 volts. Thus resulted a series of curves for the various bias potentials used experimentally. Other values, and other tubes, could be used to obtain the same fundamental results. The plate of the tube never goes negative because of the fact that the d-c potential acting in the plate circuit is larger than the signal voltage. It is also observed that the grid of the tube is not allowed to go positive, but for one extreme value the grid is at zero potential with respect to ground. It is always negative with respect to its cathode, however, for even though no bias battery potential is used, there is a d-c voltage drop in the cathode circuit resistor, which acts as an automatic bias potential. The polarity of this voltage is indicated in the figure.

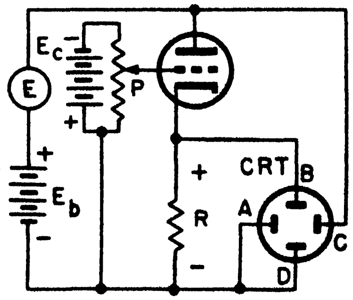


Fig. 19-23. — Circuit setup used for obtaining characteristic curves of Fig. 19-22.

Grid Potential Changes

Rather than a variation in the plate signal potential, we may show, by means of Fig. 19-24A, the changes in plate current that occur during the signal cycle in the grid circuit. The circuit arrangement is shown in Fig. 19-24B. For any particular curve, the plate potential

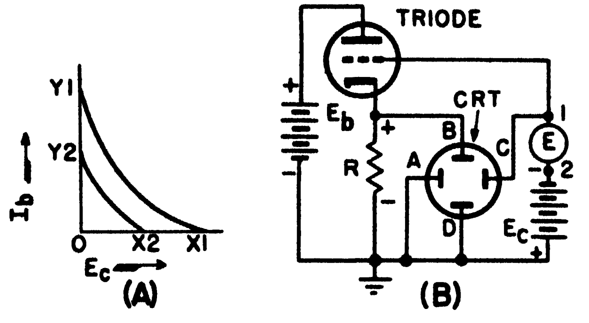


Fig. 19-24.—(A) Changes in plate current as a function of grid signal voltage; (B) circuit for checking performance of tube under dynamic signal conditions.

and d-c bias voltages are held constant. Suppose that the highest permissible plate potential is used. Then, the maximum value of plate current is *Y1*. Similarly, using a lower value, the maximum is *Y2*. The bent part of the curve illustrates the nonlinearity of the tube characteristic. The curve is usually reasonably flat over a major portion of its length. It is over this flat portion that operation is often sought in the design of circuits using such tubes. For detection, or mixing, the nonlinear characteristic is not only desirable but necessary. A photograph of a series of curves obtained on a cathode-ray-tube screen is shown in Fig. 19-25. The signal voltage *E* had an rms value of 8.86 volts and the bias potential was set at −12.5 volts.

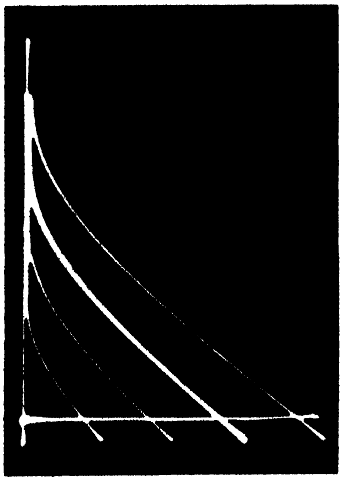


Fig. 19-25.—Series of curves obtained on cathode-ray-tube screen for triode class A operation.

The grid is not allowed to go positive during the signal cycle, which is the condition required for the usual Class A functioning of an amplifier tube. Various values of plate potential, 450, 350, 250 and 150, were used. As the grid never goes positive with respect to ground, plate *C* of the cathode-ray tube is always at a negative potential and repels electrons of the beam within the cathode-ray tube away from this plate. The motion is toward the left due to this force. The motion of the beam is also upward due to a force caused by the positive, attractive charge on *B* of the cathode-ray tube. The net resultant force is the vector sum of the two described forces. The voltage across *R* is in phase with *E*. Therefore, as *E* becomes positive, at terminal *1* with respect to terminal *2*, the bias is reduced on the grid and the cathode current through *R* goes up, the voltage across this resistor being in phase with the current. The shape of the curve is shown in Fig. 19-24A and the family of curves in Fig. 19-25.

The dynamic characteristic of a tube under specified conditions is similar to the characteristic curves just obtained, except that it takes into consideration the reduction in voltage, at the plate of the tube, due to the load impedance. (An adequate description of tube characteristics and their meaning will be found in a good radio tube handbook.)

The dynamic characteristic curve, in general, will not be a straight line but will have a slight curvature. The amount of this curvature is decreased by increasing the amount of resistance used as a plate-circuit load.

Making the assumption that the curve, for purposes of analysis, is parabolic rather than a straight line, it can be shown mathematically that the application of a sine-wave input signal to the grid will result in a plate current that contains a second-harmonic term and constant term in addition to the fundamental frequency. The constant term arises from the partial rectification that takes place as a result of the curvature in the dynamic curve of the tube. However, if the plate-load circuit is reactive, the curve becomes an ellipse. This is illustrated in Fig. 19-26. It is known that the equivalent circuit analysis often used for making engineering calculations cannot accurately be applied in this case. The shapes and general characteristics of the curves are dependent on the circuit arrangement used to take them.

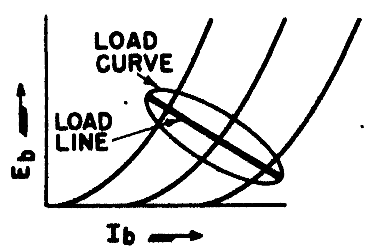


Fig. 19-26.—Effect of reactive plate load is that the load curve takes on the shape of an ellipse.

Tetrodes

An oscillogram showing the plate, screen, and cathode currents through a tetrode is shown in Fig. 19-27. The total cathode current is practically constant. These relative effects are shown by the graphical portrayal in the figure. The screen current is lower than that of the plate since it generally operates at a lower potential.

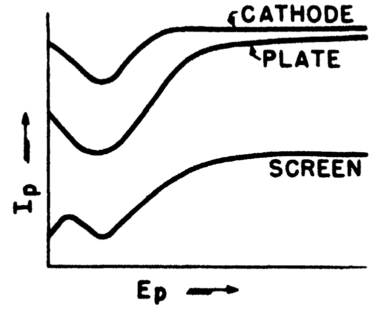


Fig. 19-27.—Oscillogram showing plate, screen, and cathode currents through a tetrode.

Pentodes

The pentode-type tube is widely used as a voltage and power amplifier. A means of checking pentode-type tubes is illustrated in Fig. 19-28. The vertical-deflection plates of the oscilloscope are connected across resistor R, which is in series with the negative line of the B potential source. E is a signal source, such as an audio generator, which supplies an alternating potential to the grid of the pentode under test. E_s is the screen potential.

Sharp-cutoff pentode voltage-amplifier tubes are used, sometimes as detectors or oscillators. Remote-cutoff pentode voltage-amplifier tubes are customarily used as r-f or i-f amplifiers, and tubes with intermediate or remote characteristics may be used as audio voltage amplifiers or power output tubes. Curves showing the various types of functions are shown in Fig. 19-29.

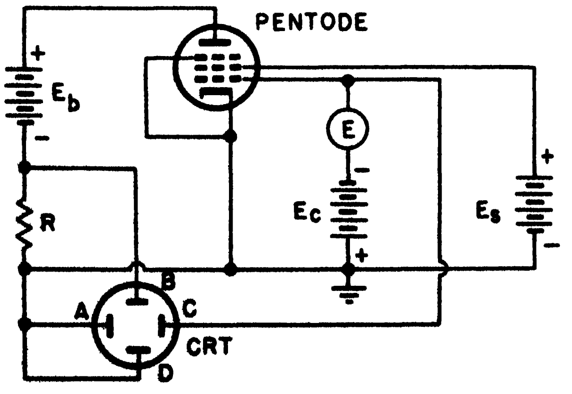
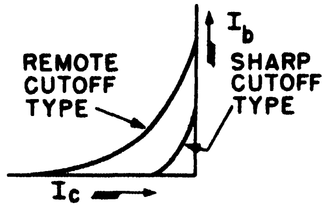


Fig. 19-28.—Circuit for checking pentode tube characteristics with cathode-ray tube.

Fig. 19-29.—Types of pentode characteristic curves; sharp and remote cutoff.



Beam Power Tubes

The beam power tube is an improvement over the pentode type, having greater dynamic power sensitivity. Its constructional features are described in standard texts and tube manuals. Oscillograms of the tube characteristics of a typical beam type are shown in Fig. 19-30, for various values of grid potential.

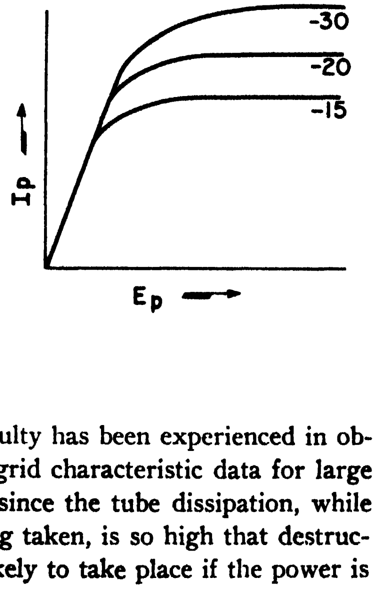


Fig. 19-30.—Typical beam power tube characteristic curves.

Positive Grid Tests[1]

Considerable difficulty has been experienced in obtaining the positive grid characteristic data for large high-vacuum tubes, since the tube dissipation, while these points are being taken, is so high that destruction of the tube is likely to take place if the power is left on sufficiently long to read a conventional meter. To obviate this difficulty, various methods have been devised. In general, these consist of means for applying proper voltages to the tube under test, for very short times, and for recording photographically the voltages and currents seen oscillographically. In the case of very large tubes that draw high currents, a technique that may be used requires a rather large

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[1] The following material is due to O. W. Livingston, Proceedings of I.R.E., vol. 28, no. 6, p. 269, June, 1940.

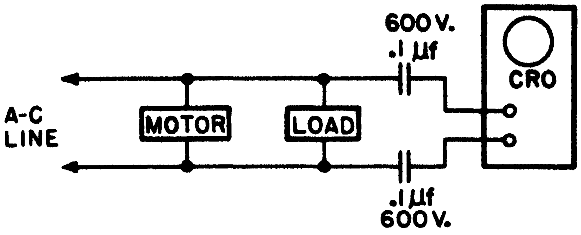


Fig. 19-35.—Circuit for checking a-c line voltage waveform.

under consideration. When the motor is switched on and off, the line voltage will be affected. Unless the oscilloscope has a built-in electronic voltage regulator, the trace on the cathode-ray-tube screen may wobble quite a bit and the operation of the equipment under test will also be affected. This equipment is represented as a load in the drawing. The motor might be replaced by any other electrical device drawing a good deal of power from the line, thereby affecting the voltage regulation. It is important to keep in mind, in making tests, the conditions under which they are made, so that true interpretation of the patterns observed on the cathode-ray-tube screen may be made. If, for example, a constant-voltage transformer is used according to Fig. 19-36, to stabilize the source potential, it will be necessary to take into account the slight distortion of the line waveform, or, rather, the waveform of the voltage delivered by the transformer to the load. A departure from a sine wave may be expected. The filter waveforms, however, in half-wave and full-wave rectifier systems, will be so slightly affected that the result may be regarded as negligible.

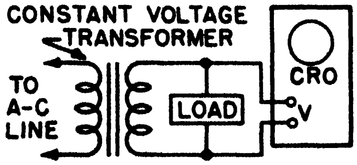


Fig. 19-36.—Circuit for checking waveform across secondary of a constant voltage transformer.

Power Supplies

The function of a power supply is simply to supply power to the circuit. In a battery set, the battery system represents the simplest form of power supply. In the a-c or auto set, or d-c line receiver, a rectifier tube and filter system are used. The waveforms that are obtained vary according to the design and condition of the equipment. In receivers, the three main classifications are half-wave, full-wave, and vibrator-type supplies.

Half-Wave Rectifier

The basic circuit is shown in Fig. 19-37. Many commercial receivers employ a half-wave rectifier of this type. High-voltage capacitors were used in this experimental setup to avoid damage that might result during tests. Two oscilloscopes were used for the tests; a Waterman S11-A was used for usual observation and an RCA WO-79A, with attached camera, permitted photographing the various traces reproduced in this chapter.

R_L consisted of ten 300-ohm, 10-watt resistors in series. A shorting wire (insulated except for the ends), fitted with alligator clips, permitted varying the load on the supply. An input of 115 volts was applied to the rectifier circuit through a standard General Radio Variac and held at that value by referring to a 0-150 volt a-c voltmeter shunted across the rectifier a-c input circuit.

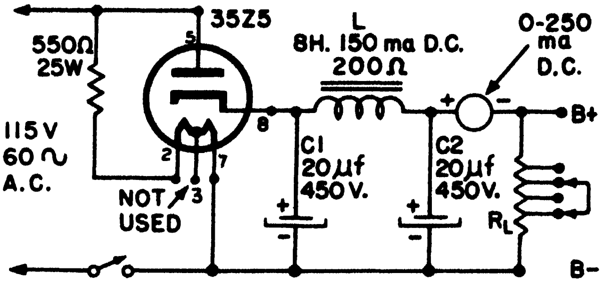


Fig. 19-37.—Basic half-wave rectifier circuit.

The waveform obtained with the oscilloscope input connected across *C1*, the input filter capacitor, Fig. 19-37, is shown in Fig. 19-38. The value of R_L was adjusted to draw a current of 0.04 ampere (40 ma).

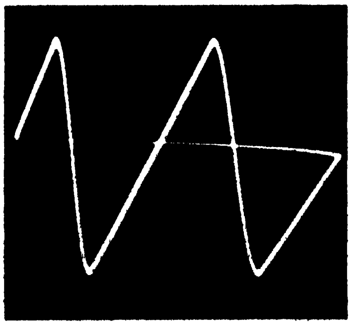


Fig. 19-38. — Waveform obtained with the oscilloscope input connected across *C1*, the input filter capacitor of Fig. 19-37.

The d-c voltage across *C1* was 135 volts and that across *C2* was 128 volts. Without changing the vertical gain and still using "line sync," the pattern in Fig. 19-39 was obtained, which shows practically a pure d-c output across *C2*.

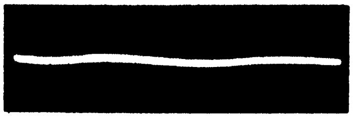


Fig. 19-39. — Waveform of voltage output across *C2*.

The pattern in Fig. 19-40 was obtained using the oscilloscope across *C2* with inductance *L*, Fig. 19-37, shorted out of the circuit. This is also the waveform for *C1*, since, with *L* shorted out, *C2* is in parallel with *C1*. With no load on *C1* (R_L and the choke disconnected) a practically straight line pattern, as shown in Fig. 19-41, was obtained. With R_L adjusted for a 90-ma load (heavy loading of the power supply) the a-c voltage component across *C1* increased in magnitude, as shown by the pattern in Fig. 19-42.

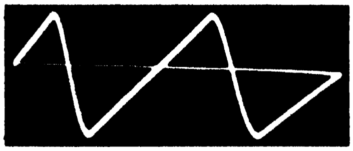


Fig. 19-40. — Waveform of voltage across *C2* with industance *L* (Fig. 19-37) shorted out of the circuit.

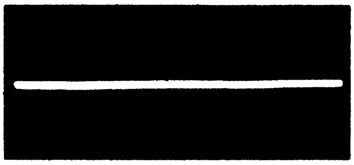


Fig. 19-41.—With no load on *C1* (R_L and the choke disconnected) a practically straight-line pattern is obtained.

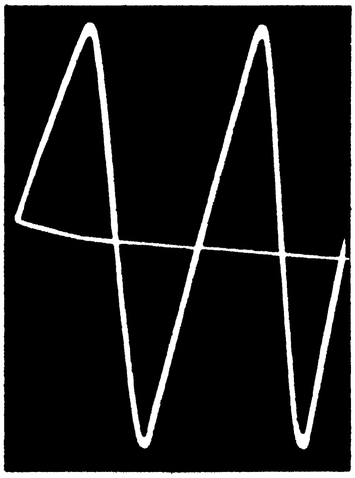


Fig. 19-42. — Heavy loading of the power supply produced this pattern.

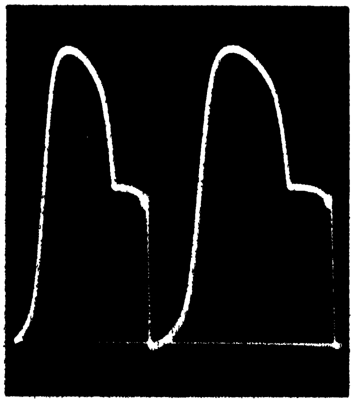


Fig. 19-43. — Waveform of voltage between rectifier cathode and B— with *C1* out of circuit.

Some rather interesting patterns are obtained under various conditions. Using $R_L = 3{,}000$ ohms, and removing *C1* from the circuit, the waveform of the voltage between the rectifier cathode and $B-$ was that shown in Fig. 19-43. A 20-ma d-c output current was obtained in R_L. With *C1* open, the waveform of the voltage across *L* was that shown in Fig. 19-44.

The waveform of the a-c voltage across *C1*, with *C2* open, is shown in Fig. 19-45. With *C2*, some ripple was observed across R_L, as shown by Fig. 19-46. Using the same setup as Fig. 19-37, except that a 2,000-ohm resistor *R* was substituted for *L*, different patterns, as might have been expected, were obtained. Fig. 19-47 shows the circuit. Fig. 19-48 shows the a-c voltage

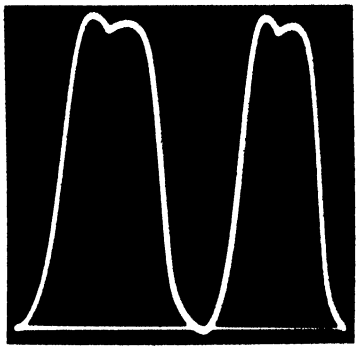


Fig. 19-44. — Waveform of voltage across *L* with *C1* removed from circuit.

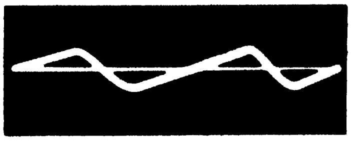


Fig. 19-45. — Waveform of voltage across normal input filter capacitor, with output filter capacitor disconnected to simulate open condition.

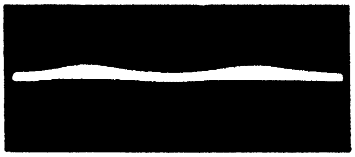


Fig. 19-46. — Waveform of voltage across output filter capacitor of half-wave supply at slight loading (20 ma), indicating small amount of hum voltage.

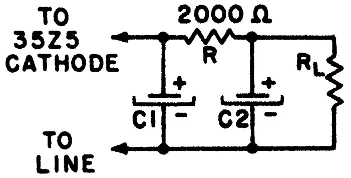


Fig. 19-47. — Circuit using series resistance (2,000 ohms) in place of filter choke.

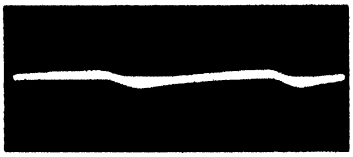


Fig. 19-48. — Waveform of voltage across input filter capacitor of *R-C* filter system, with disconnected load, in half-wave rectifier power supply.

waveform across $C1$ with R_L open. R_L was adjusted for a 50-ma load. Fig. 19-49 shows the waveform across R with $R_L = 2{,}000$ ohms and $C1$ open.

The foregoing discussion of the half-wave rectifier, using the high-vacuum diode tube is logically followed by a discussion of a similar half-wave rectifier, using a selenium element, which has found favor with receiver designers of late; typical waveforms obtained using this form of element will be shown.

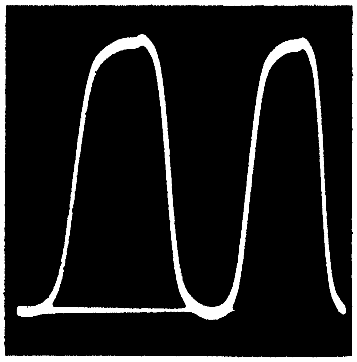


Fig. 19-49. — Waveform of voltage across series resistance in half-wave rectifier circuit with R_L set at 2,000 ohms and $C1$ open.

Selenium Rectifiers (Half-Wave)

The selenium rectifier is similar in many respects to a half-wave rectifier tube, but has a reverse as well as forward current characteristic. Fig. 19-50 shows the oscillogram of the output current of a standard selenium rectifier. Such rectifiers may have current ratings of 100 to 200 ma. The resistance in series with the rectifier element varies inversely as the load current rating. Typical values are 22 to 5 ohms for the 100- to 200-ma types.

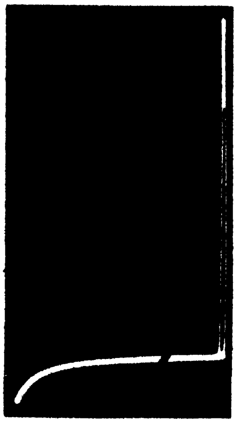


Fig. 19-50. — Oscillogram of the output current as a standard selenium rectifier.

Courtesy Radio Receptor Co.

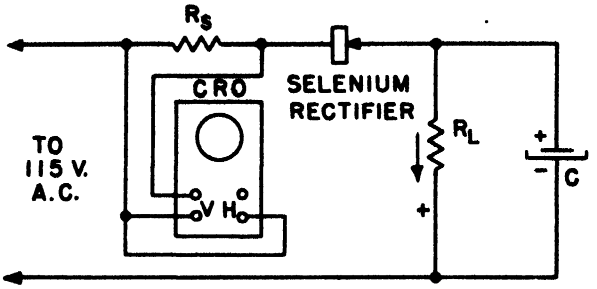


Fig. 19-51.—Circuit of selenium half-wave rectifier power supply.

In the circuit shown in Fig. 19-51, the filter capacitor C has a value of 40 μf. The load resistance is adjusted, experimentally, to draw the rated current. The internal sweep of the oscilloscope is not used, the vertical input being connected across R_s and the horizontal input across the a-c line; the oscillogram is, therefore, the curve of current plotted against voltage. Note that an external ground is not used, since it might cause a short circuit of the power line, one side of which is usually grounded.

A portion of the curve in Fig. 19-50 is practically horizontal. The reverse current graph is shown below the horizontal axis and the forward current above it. Without the capacitor in the circuit, the graph assumes the shape indicated in Fig. 19-52. Using the same circuit but omitting the horizontal input connections to the test circuit of Fig. 19-50, and using a timing axis linear sweep frequency of 30 cps in the oscilloscope, the pattern shown in Fig. 19-53 is obtained. Two relatively steep pulses are shown for the forward current characteristic and minor downward loops are indicated for the reverse current characteristic.

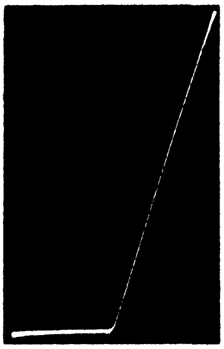


Fig. 19-52.—Oscillogram of voltage obtained when filter capacitor shunt across load resistance is not used.

Courtesy Radio Receptor Co.

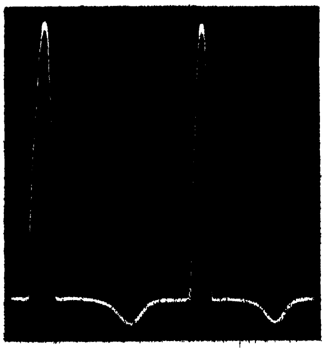


Fig. 19-53. — Pattern obtained with cathode-ray sweep set at 30 cps in half-wave 60-cps supply.

Courtesy Radio Receptor Co

A somewhat similar pattern but with broader peaks is shown in Fig. 19-54. This pattern was obtained using the oscilloscope vertical input across R_g and a 30-cps internal sweep. The oscilloscope horizontal input was disconnected from the test circuit, and the capacitor normally across R_L was disconnected so that no filter capacitance was used.

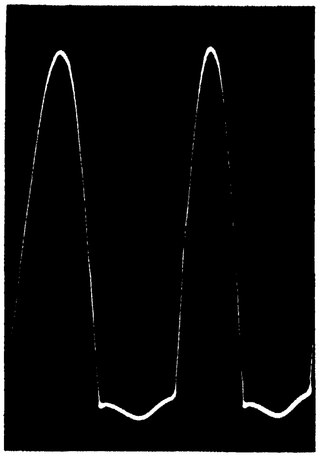


Fig. 19-54. — Oscillogram of voltage across R_L with *C* omitted using 30-cps cathode-ray oscilloscope internal sweep and no horizontal input connections to test circuit of rectifier.

Courtesy Radio Receptor Co.

Full-Wave Rectification

The waveforms obtained using full-wave rectification are somewhat different than those obtained for half-wave. The test circuit used is shown in Fig. 19-55; because of the higher voltage, a 1-megohm resistor was connected in series with the oscilloscope input circuit to reduce the oscilloscope input voltage to approximately half.

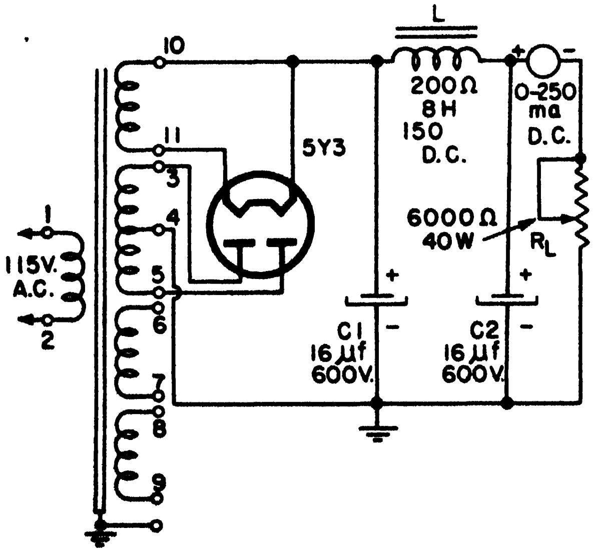


Fig. 19-55.—Full-wave rectification test circuit.

Oscillograms taken, using this circuit, are shown in Figs. 19-56 to 19-63. The details are given in the caption with each figure.

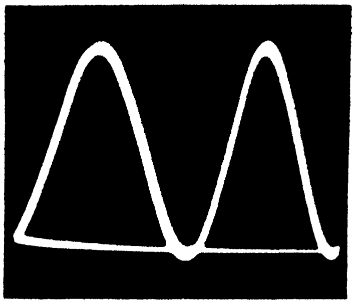


Fig. 19-56. — Waveform of voltage across primary terminals of power transformer in full-wave rectifier power-supply system.

Fig. 19-57. — Waveform of voltage across output filter capacitor *C2* for 50-ma load, using circuit in Fig. 19-55.

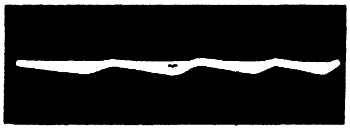


Fig. 19-58. — Voltage across *L* (Fig. 19-55) for 40-ma load.

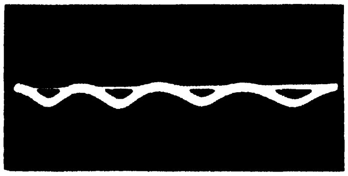


Fig. 19-59. — Waveform of voltage across output filter capacitor *C2* of full-wave supply with input filter capacitor *C1* disconnected, circuit of Fig. 19-55.

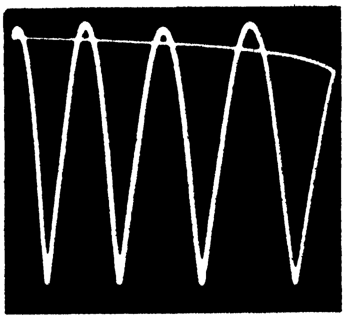


Fig. 19-60. — Waveform of voltage from input of *L* to ground, capacitor *C1* disconnected, circuit of Fig. 19-55.

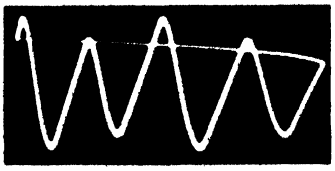


Fig. 19-61. — Waveform from input of *L* to ground, *C1* input filter capacitor connected, *C2* output filter capacitor open.

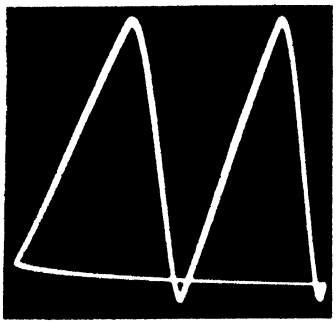


Fig. 19-62.—Waveform of voltage across *L*, series filter inductance, with one rectifier plate circuit open, circuit of Fig. 19-55.

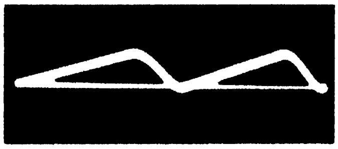


Fig. 19-63. — Voltage across the input filter capacitor with one rectifier plate circuit open, circuit of Fig. 19-55.

Selenium Rectifier (Full-Wave)

The selenium rectifier is widely used commercially in industrial equipment, particularly in bridge circuits such as that in Fig. 19-64. The normal pattern of the ripple voltage across the load resistance, R_L, is shown in Fig. 19-65 which was obtained using a 30-cps sweep. This pattern is also obtained when the circuit of Fig. 19-66 is examined.

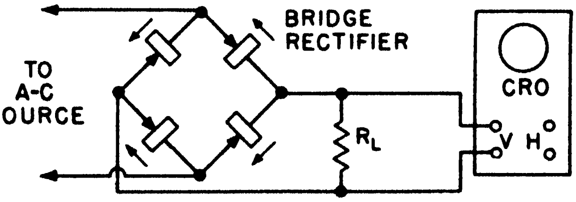


Fig. 19-64.—Test setup for determining bridge-rectifier characteristics. Arrows show electron current flow.

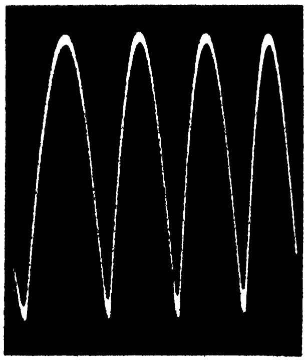


Fig. 19-65. — Normal ripple voltage across load. Sweep at half of power line frequency, circuit of Fig. 19-64.

Courtesy Radio Receptor Co.

The pattern obtained when any one arm of either type circuit is open consists of two pulses, and is the same as the pattern shown, earlier in the selenium half-wave rectifier discussion, for the condition when a 30-cps internal oscilloscope sweep is used and the capacitor across the load is omitted.

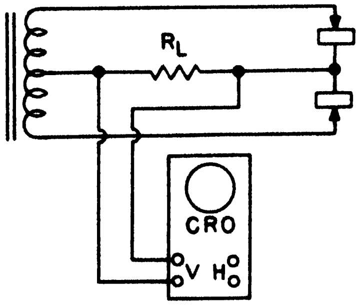


Fig 19-66 —Test setup for obtaining characteristics of full-wave rectifier.

The pattern in Fig. 19-67 was obtained for the ripple voltage across R_L with a circuit condition such that the rectifiers had greatly unmatched forward resistances. This condition may be observed when one arm has been damaged.

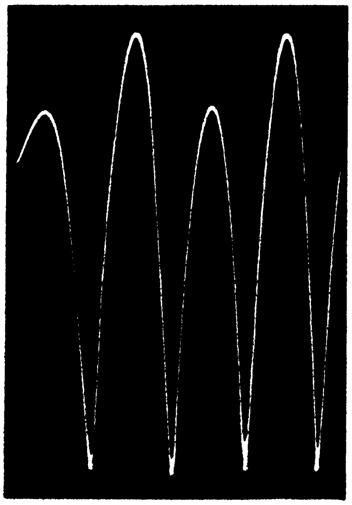


Fig. 19-67.—Ripple voltage with any one arm open, circuit of either Fig. 19-64 or Fig. 19-66.

Courtesy Radio Receptor Co.

Three-Phase Rectifiers

This type of circuit is used in industrial apparatus. Shown in Fig. 19-68, *L1, L2,* and *L3* are the transformer secondaries. The various selenium rectifier elements are designated *x1* to *x6*. R_L is the load resistance. The oscillograms were taken using a 30-cps

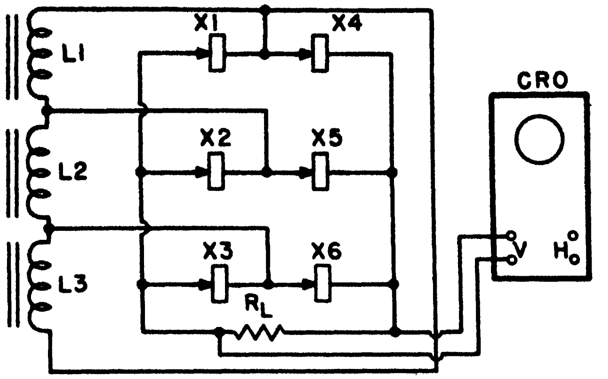
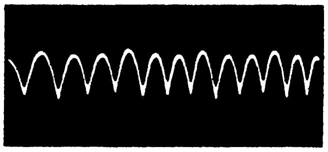


Fig. 19-68.—Test circuit setup for industrial three-phase bridge rectifier.

sweep for 60-cps line frequency. Figs. 19-69 to 19-72 are oscillograms taken across a three-phase rectifier load, as in Fig. 19-68, under conditions stated in the captions.

Fig. 19-69. — Normal three-phase ripple voltage across load, circuit of Fig. 19-68.



Courtesy Radio Receptor Co.

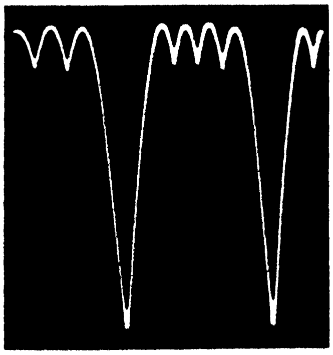


Fig. 19-70.—Ripple voltage across load with one arm of circuit open, circuit of Fig. 19-68.

Courtesy Radio Receptor Co.

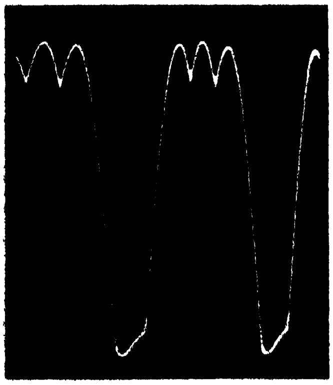


Fig. 19-71.—Ripple across load with one negative and one positive arm (not paired) open, circuit of Fig. 19-68.

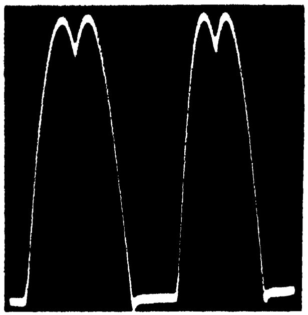


Fig. 19-72.—Ripple voltage across load with two negative or two positive arms open, circuit of Fig. 19-68.

Courtesy Radio Receptor Co.

Vibrator Testing

To the average radioman, a vibrator means just one thing — a part of an auto radio power-supply system. However, largely due to war research, vibrator applications have been increased, until now virtually every field of electronics using d.c. as a primary source of power and needing an a-c power circuit uses, in one form or another, a vibrator-type mechanism. The same principles which apply to auto vibrator design and usage also apply to vibrators in other equipment. Of course, it is necessary to correlate other factors such as power rating, continuous duty performance, etc.

The photographs of voltage waveforms obtained under various conditions that are shown in this chapter are general, not specific illustrations. A certain amount of this information is basic and can be applied without great discrimination, but there are too many various, too many complex factors, to permit breaking down the patterns into groups that would apply to specific problems. As an analogy, the general classes of fingerprints may be organized, and rules established for identification, but the fingerprints of one individual are never exactly like those of another person.

The proper adjustment of the contacts entails the consideration of long life, as well as operating efficiency. Both of these factors, efficiency and long life, are related to specific modes of application with respect to the related apparatus, such as the power transformer, the load, buffer capacitors, etc.

Speaking generally, symmetry in the voltage waveform is an essential factor. Comparative freedom from arcing, which is productive of noise and appears in the oscilloscope image as jagged streamers or off-shoots, is required. Imperfect operation, due to one cause or another, produces incorrect output voltage and incorrect primary current. Usually, faulty vibrator operation results in lowered B-supply output voltage and increased battery current. Correction of the condition restores the output voltage to its normal value. Correlation of the oscilloscope pattern and a measurement of the d-c output voltage serves as a guide to proper adjustment of the vibrator. A 5,000-ohm load is usually used when testing auto radio vibrator units, as this is an average value.

Speaking of symmetry in the voltage waveform, a slight departure is to be expected when checking the operation of that portion of the transformer windings which is related to the starting winding. This winding shunts one-half of the primary winding in all full-wave, synchronous, and nonsynchronous vibrator systems. The oscillograms to be shown will serve to illustrate the applications of the cathode-ray oscilloscope in the interpretation of electronic circuit observations obtained visually. Information of this kind is useful to the serviceman. The schematic diagram of the full-wave nonsynchronous vibrator power-supply unit used during some of these tests is shown in Fig. 19-73.

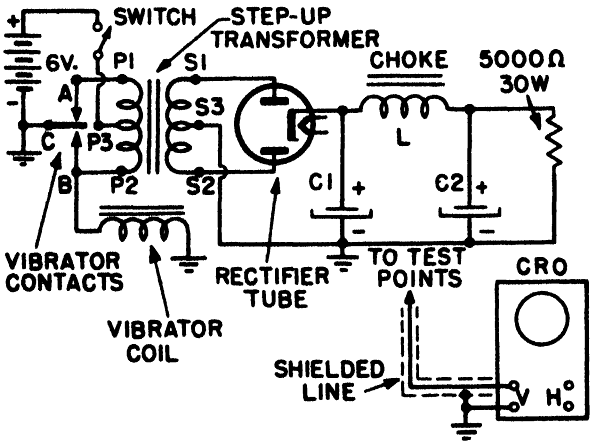


Fig. 19-73.—Schematic diagram of full-wave nonsynchronous vibrator power-supply unit used in tests.

The 6-volt battery may be any standard automobile battery. Usually, the negative side is grounded but the polarity is unimportant for a nonsynchronous vibrator. For the synchronous type, observation of proper polarity is essential. The rectifier tube may be any standard type. *C1* and *C2* may be 4- to 20-μf 450-volt capacitors, *L* is a choke of 5 to 20 henrys, R_L is a 5,000-ohm, 30-watt wire-wound resistor.

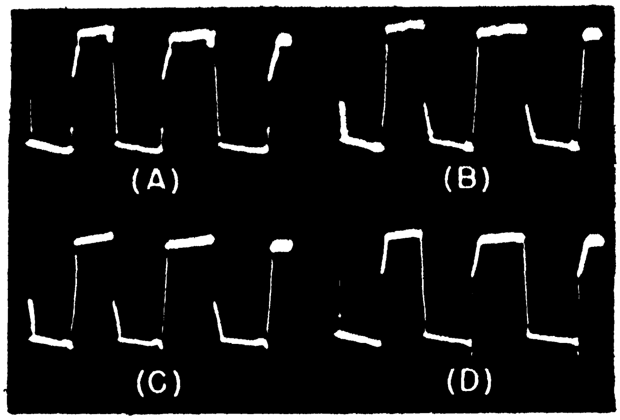


Fig. 19-74.—Oscillograms taken at transformer, Fig. 19-73. (A) and (B) show primary voltages, *P1-P3*, *P2-P3*. (C) and (D) show secondary voltages, *S1-S3*, *S2-S3*.

The oscilloscope is connected as shown in Fig. 19-73, the *Y*-axis input going to the B-supply output circuit, or other test point. The oscilloscope was used as a means of observing voltage waveforms. The internal linear sweep was used, with internal synchronization. Vibrator adjustments were checked by noting the waveforms of the various voltages. Oscillograms of voltages taken across *P1-P3* and *P2-P3* are shown in Fig. 19-74A and Fig. 19-74B, respectively. Oscillograms of voltages taken across *S1-S3* and *S2-S3*, respectively, are shown in Fig. 19-74C and Fig. 19-74D. An oscillogram of the voltage across *C1* is shown in Fig. 19-75. The oscillograms in Figs. 19-76A, 19-76B,

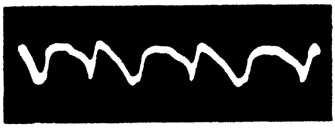


Fig. 19-75.—Oscillogram of voltage across input filter capacitor of vibrator-type power supply, circuit of Fig. 19-73.

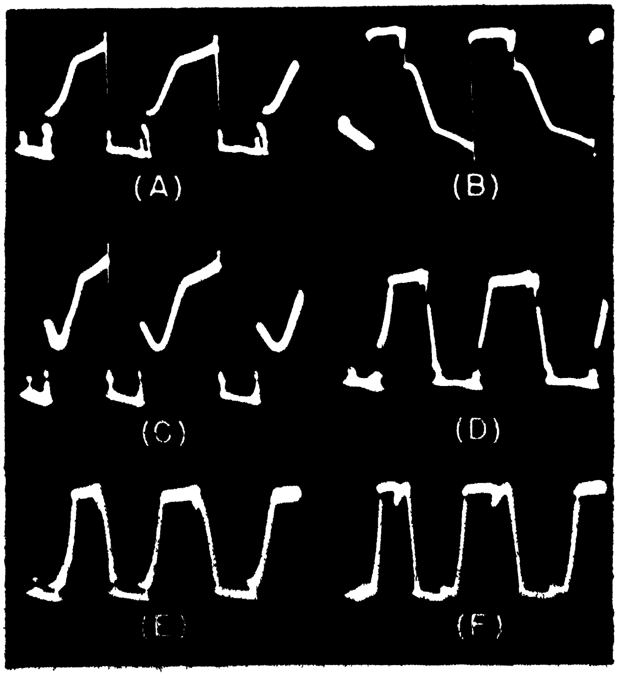


Fig. 19-76.—Vibrator circuit waveforms. (A), (B), and (C) show half-wave vibrator operation, contact *A* (Fig. 19-73) disconnected; parts (D), (E), and (F) show full-wave operation under certain conditions.

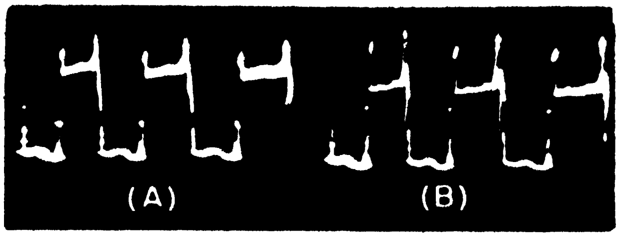


Fig. 19-77.—Effect of unbalanced input circuit (refer to Fig. 19-73). (A) resistance of 1 ohm in series with contact *A*; (B) resistance of 2 ohms in series with contact *A*.

and 19-76C were made with the vibrator of Fig. 19-73 functioning as a half-wave vibrator, contact *A* being disconnected. The *P1-P3* voltage waveform is shown in Fig. 19-76A. That across *S1-S3,* also using half-wave vibrator operation, is shown in Fig. 19-76B.

Fig. 19-76C shows the *P1-P3* voltage when the remaining active contact clearance or separation of the vibrator was increased appreciably. Fig. 19-76D shows the *P1-P3* voltage with vibrator contact *A* reconnected normally. The contact separation was made uniform so that symmetry was obtained. Fig. 19-76E shows the *P1-P3* voltage, starting with the adjustment just described and then opening the contacts to maximum operation. The voltage across *S1-S3* is shown in Fig. 19-76F. Fig. 19-77 A shows the pattern obtained for a 1-ohm resistance in series with contact *A* of the vibrator, and Fig. 19-77B shows the waveform with 2 ohms added to the *A* vibrator circuit.

Full-Wave Synchronous

The basic circuit is shown in Fig. 19-78, and it is, of course, at once obvious that this circuit is distinguished from the nonsynchronous by the absence of a rectifier tube. The vibrator itself does the rectifying, rather than an electronic rectifier.

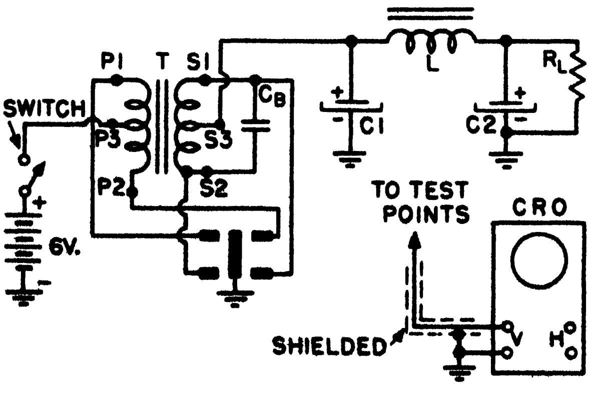


Fig. 19-78.—Schematic diagram of full-wave synchronous vibrator test arrangement.

With the supply in normal working order, the voltage waveforms in Fig. 19-79A and B, were obtained, across *P1-P3* and *P2-P3* respectively. The corresponding secondary voltages are Figs. 19-79C and 19-79D. Fig. 19-79E shows the waveform of the voltage across *C1,* the input filter capacitor. Fig. 19-79F is a duplicate of Fig. 19-79E except for two items: Fig. 19-79E is a 3-second exposure whereas Fig. 19-79F is a ½-second exposure. A transient is present in Fig. 19-79F, but the "streamers" or "off-shoots," which cause r-f interference, although visible upon the screen, did not record upon the film. The long exposure is the reason for the very thick lines in Fig. 19-79E.

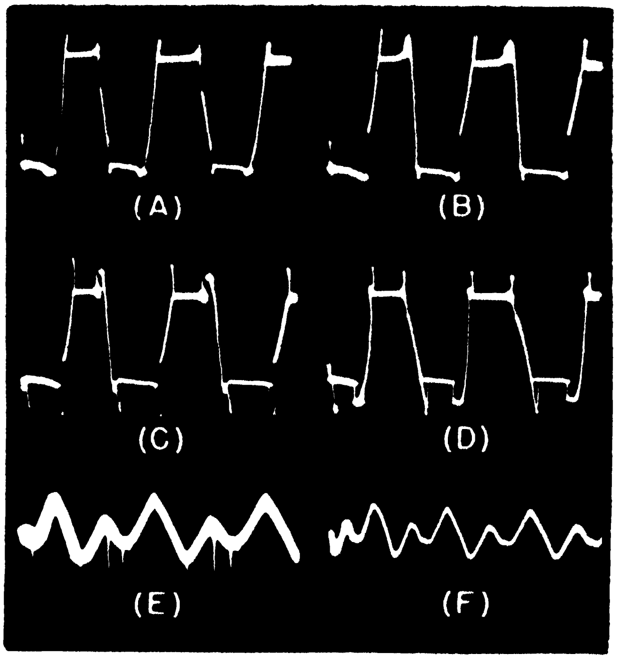


Fig. 19-79.—Oscillograms of normal operation of synchronous vibrator (refer to Fig. 19-78): (A) primary voltage *P1-P3;* (B) primary voltage *P2-P3;* (C) secondary voltage *S1-S3;* (D) secondary voltage *S2-S3;* (E) voltage at filter input across *C1.* Note fine ***streamers*** caused by r-f interference, brought out by long photographic exposure, which also thickens main trace; (F) same as (E), except normal exposure. Streamers are not recorded photographically, although visible to the eye.

Maximum separation between the secondary contacts resulted in the waveform of voltage across *P1-P3* shown in Fig. 19-80A. Arcing was observed at the contacts. The effect of very little spacing between the contacts is shown in Fig. 19-80B. The voltage was that across *P1-P3.* A readjustment for symmetrical waveform developed the pattern shown in Fig. 19-80C.

of the d-c output voltage serves as a guide to proper adjustment of the vibrator. A 5,000-ohm load is usually used when testing auto radio vibrator units, as this is an average value.

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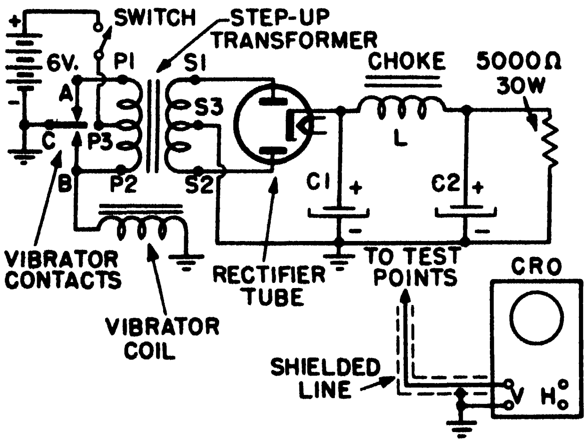

Fig. 19-73.—Schematic diagram of full-wave nonsynchronous vibrator power-supply unit used in tests.

The 6-volt battery may be any standard automobile battery. Usually, the negative side is grounded but the polarity is unimportant for a nonsynchronous vibrator. For the synchronous type, observation of proper polarity is essential. The rectifier tube may be any standard type. *C1* and *C2* may be 4- to 20-μf 450-volt capacitors, *L* is a choke of 5 to 20 henrys, R_L is a 5,000-ohm, 30-watt wire-wound resistor.

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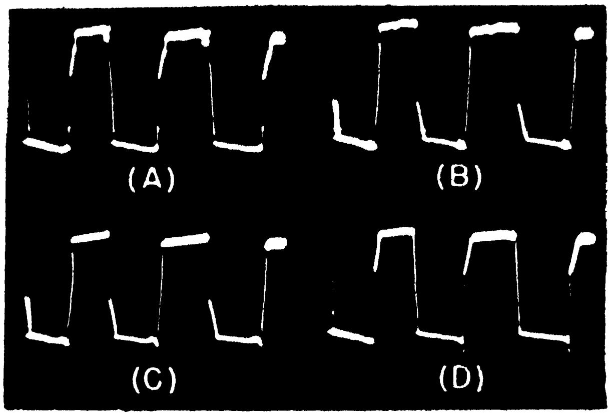


Fig. 19-74.—Oscillograms taken at transformer, Fig. 19-73. (A) and (B) show primary voltages, *P1-P3, P2-P3.* (C) and (D) show secondary voltages, *S1-S3, S2-S3.*

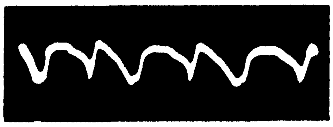


Fig. 19-75.—Oscillogram of voltage across input filter capacitor of vibrator-type power supply, circuit of Fig. 19-73.

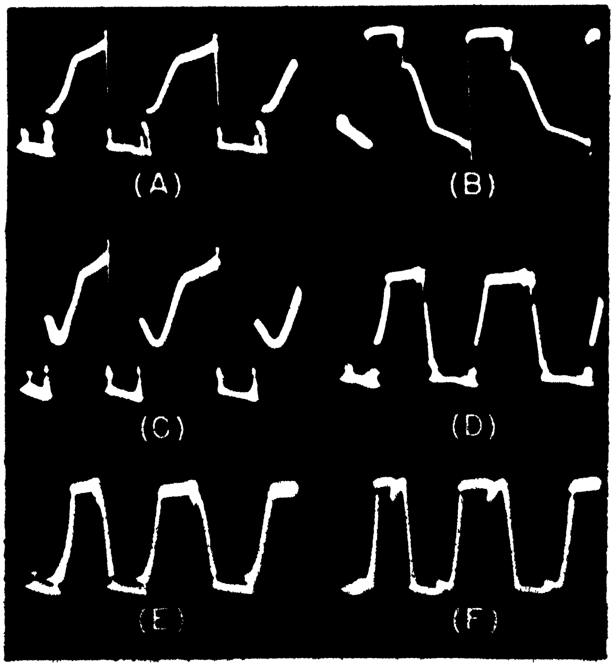
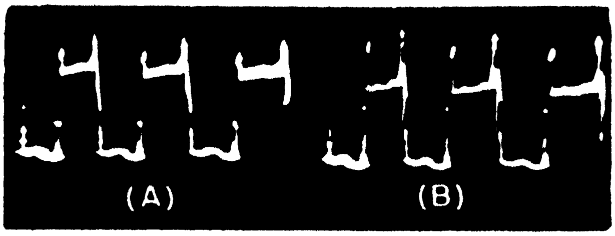


Fig. 19-76.—Vibrator circuit waveforms. (A), (B), and (C) show half-wave vibrator operation, contact *A* (Fig. 19-73) disconnected; parts (D), (E), and (F) show full-wave operation under certain conditions.



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Full-Wave Synchronous

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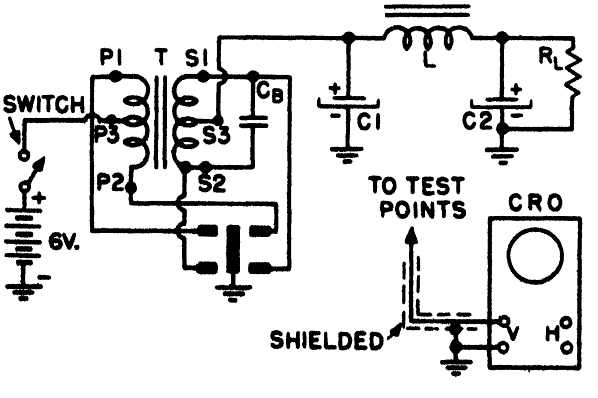


Fig. 19-78.—Schematic diagram of full-wave synchronous vibrator test arrangement.

With the supply in normal working order, the voltage waveforms in Fig. 19-79A and B, were obtained, across *P1-P3* and *P2-P3* respectively. The corresponding secondary voltages are Figs. 19-79C and 19-79D. Fig. 19-79E shows the waveform of the voltage across *C1,* the input filter capacitor. Fig. 19-79F is a duplicate of Fig. 19-79E except for two items: Fig. 19-79E is a 3-second exposure whereas Fig. 19-79F is a ½-second exposure. A transient is present in Fig. 19-79F, but the "streamers" or "off-shoots," which cause r-f interference, although visible upon the screen, did not record upon the film. The long exposure is the reason for the very thick lines in Fig. 19-79E.

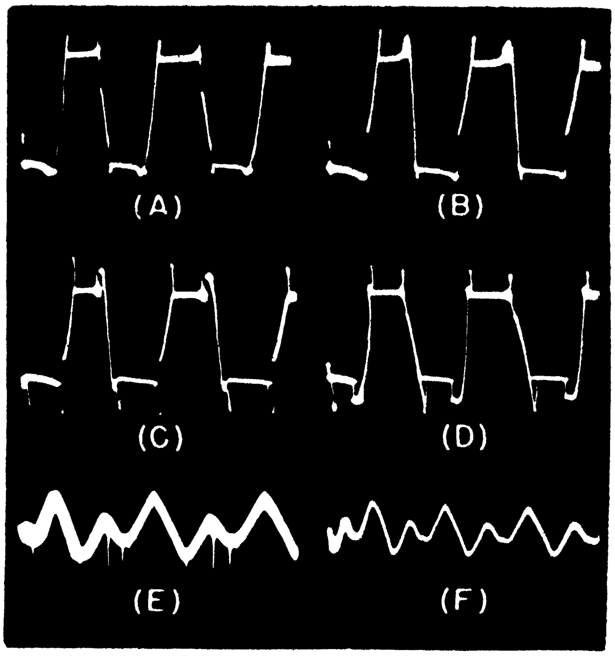


Fig. 19-79.—Oscillograms of normal operation of synchronous vibrator (refer to Fig. 19-78): (A) primary voltage *P1-P3;* (B) primary voltage *P2-P3;* (C) secondary voltage *S1-S3;* (D) secondary voltage *S2-S3;* (E) voltage at filter input across *C1.* Note fine *streamers* caused by r-f interference, brought out by long photographic exposure, which also thickens main trace; (F) same as (E), except normal exposure. Streamers are not recorded photographically, although visible to the eye.

Maximum separation between the secondary contacts resulted in the waveform of voltage across *P1-P3* shown in Fig. 19-80A. Arcing was observed at the contacts. The effect of very little spacing between the contacts is shown in Fig. 19-80B. The voltage was that across *P1-P3.* A readjustment for symmetrical waveform developed the pattern shown in Fig. 19-80C.

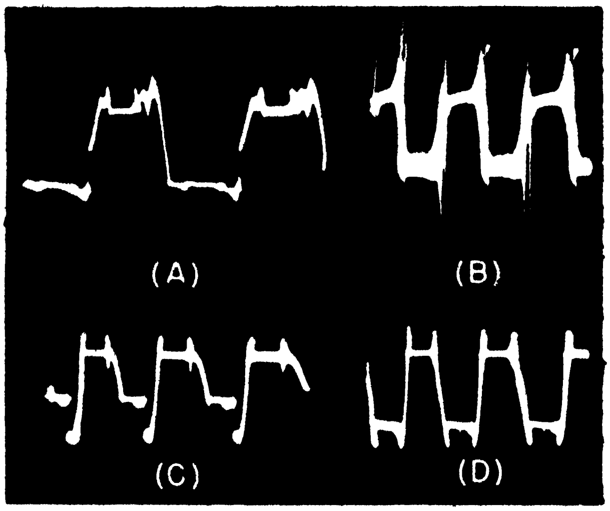


Fig. 19-80.—Effects of variation of contact spacing. Waveforms show voltage across *P1-P3*, circuit of Fig. 19-78, for various conditions.

Connecting a 0.5-ohm resistor in series with the battery produced the oscillogram of Fig. 19-80D. The B-supply output voltage also dropped from a normal of about 245 volts to about 150 volts. This is the d-c voltage measured across R_L in Fig. 19-78.

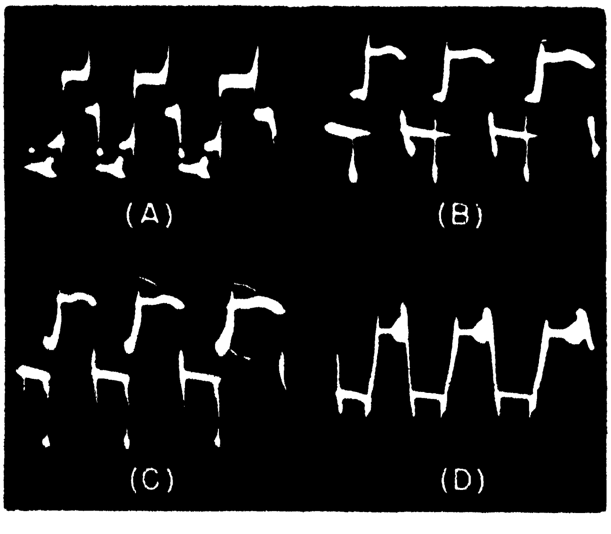


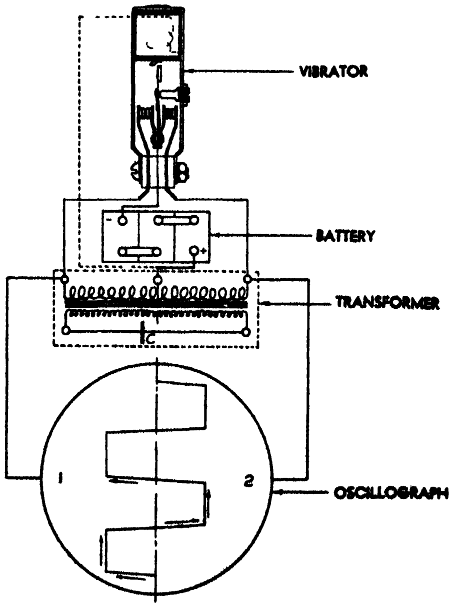
Fig. 19-81.—Effect of varying buffer capacitor C_B, circuit of Fig. 19-78, all contact adjustments normal.

Starting with the normal adjustment of Fig. 19-80, removal of the buffer capacitor C_B developed the primary voltage waveform shown in Fig. 19-81A. The waveform of the voltage in the secondary circuit is shown in Fig. 19-81B with C_B out of the circuit. The excessive amount of sparking is clearly indicated by the oscillogram. With a 0.001-μf buffer capacitor, the voltage rose to 170 volts, and the voltage across the secondary winding is shown in Fig. 19-81C. With 0.005 μf as the buffer capacitor, the voltage rose to 244 volts, with the waveform shown in Fig. 19-81D. In modern auto sets, the buffer capacitance may be approximately 0.006 to 0.01 μf for many vibrator circuits.

The oscillograms shown are indicative only of general facts and are not intended as patterns for comparison with waveforms obtained under different conditions and with other circuits.

Production of Vibrator Tests[1]

In the production of millions of vibrators used in radio equipment and electronic apparatus, the oscilloscope has been an invaluable instrument and its value increases constantly as new uses for it are found. Basically, the vibrator is a high-speed reversing switch used to reverse the direction of current flow through an associated transformer winding. By means of the vibrator, conversion of d.c. to a.c. is made possible. The most comprehensive electrical characteristic of vibrator performance is the waveform of the voltage across its two pairs of primary contact points while operating with normal load, transformer, buffer, and generally



Courtesy Du Mont Labs.

Fig. 19-82.—Production test setup for nonsynchronous vibrators.

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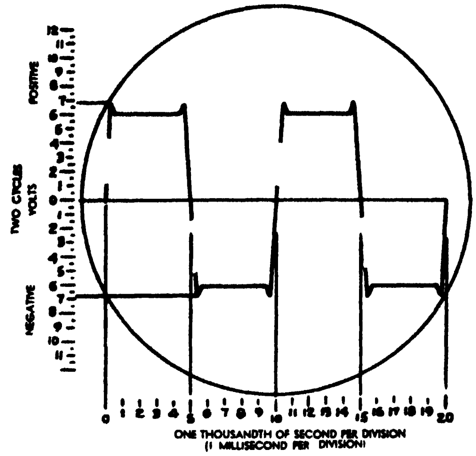
[1] "The Oscillograph as a production tool," *Du Mont Oscillographer*, vol. 6, pp. 1-8, July, August, September, October, 1943; the following material is based on this article.

normal circuit characteristics. The oscilloscope makes such waveforms conveniently visible. From the waveform, at least seven important electrical characteristics of vibrator adjustment may be judged quickly from one composite observation.

Fig. 19-82 indicates, in simplified form, the connection of the oscilloscope to the vibrator. In this drawing, we associate the upper half-cycle with the No. *1* contacts on the vibrator, and the lower with the No. *2* contacts.



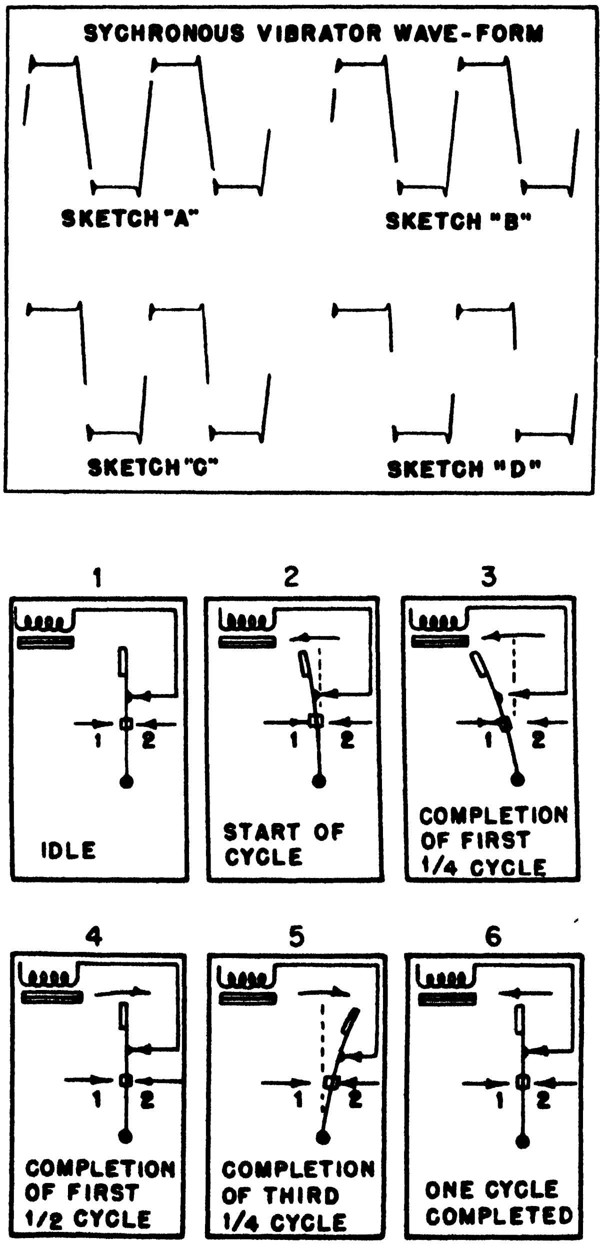
Courtesy Du Mont Labs.

Fig. 19-83.—100-cps synchronous vibrator waveform pattern.

Fig. 19-83 shows a 100-cps synchronous vibrator waveform pattern. The contact closure time for the primary is 43 per cent; for the secondary, 40 per cent. The waveform appears on its normal horizontal axis of the cathode-ray-tube screen. The pattern shows the reversal of current flow accomplished by the vibrator. The oscilloscope, in this example, essentially is an a-c voltmeter, with zero center on the horizontal baseline and with positive values above it and negative below it.

Closer spacing of the vibrator contact points provides a higher closure time and less open time during the cycle; contacts spaced wider provide a lower contact closure time. This fact may be noted by observation of the waveform. The waveform illustrated in Fig. 19-83, with a 43 per cent primary contact closure and 40 per cent secondary contact closure, indicates that the horizontal portion of each half-cycle is 40 per cent of the total cycle. This represents the secondary closure time. The horizontal width of the flat-top portion of each half-cycle, inclusive of the peaks at the corners, is 43 per cent of the total cycle width. This represents the primary contact closure time. The difference between the two spacings is called the differential or the lag of the secondary contacts. This is represented by the peaks ("dog ears") at the corners of the horizontal lines on the waveform.

In Fig. 19-84, four different waveforms are shown. Sketch "A" illustrates comparatively wide spacings of contacts — primary 36 per cent, secondary 30 per cent. Sketch "B" illustrates waveform of contacts adjusted somewhat closer than in sketch "A" — primary 40 per

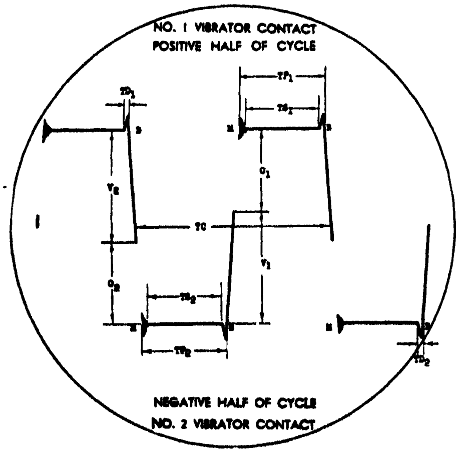

SYCHRONOUS VIBRATOR WAVE-FORM
SKETCH "A"
SKETCH "B"
SKETCH "C"
SKETCH "D"
1
2
3
IDLE
START OF CYCLE
COMPLETION OF FIRST 1/4 CYCLE
4
5
6
COMPLETION OF FIRST 1/2 CYCLE
COMPLETION OF THIRD 1/4 CYCLE
ONE CYCLE COMPLETED

Courtesy Du Mont Labs.

Fig. 19-84.—Effect of various contact spacing on waveforms. The lower drawings of this figure show one complete cycle of vibrator operation in six steps.

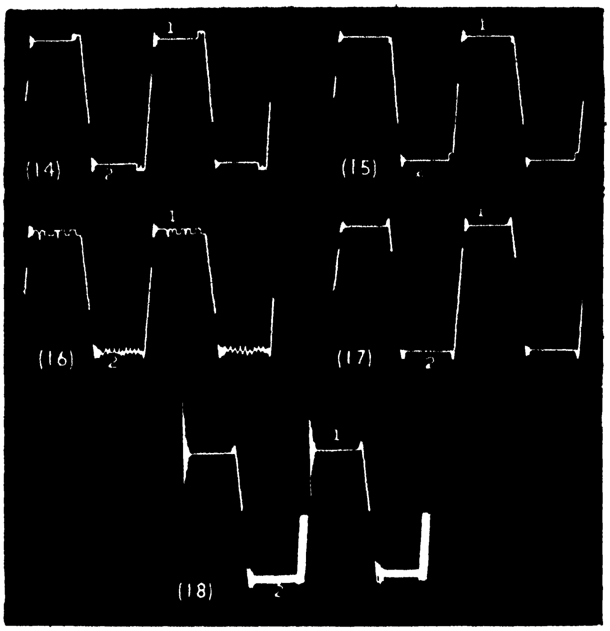
cent and secondary 34 per cent. Sketch "C" illustrates closely spaced contact points — primary 43 per cent and secondary 37 per cent. Sketch "D" illustrates very close contact spacing — primary 46 per cent and secondary 40 per cent. All four sketches indicate a differential or secondary lag of 6 per cent. This is normally the maximum lag necessary. In many applications, a differential of only 2 to 5 per cent is adequate.

It should be noted that as the contacts are spaced closer, the vertical lines of the waveform become less tilted and become nearly straight up and down on the very close-spaced adjustment, while on the wide-spaced adjustment, the vertical lines slope considerably. Another useful clue to vibrator adjustment which the oscillogram provides is the size of the opening in the vertical lines. This characteristic may be reliably used to judge contact spacing for production testing. The size of the opening is inversely proportional to the contact spacing, as can be seen from a study of the sketches in Fig. 19-84. Fig. 19-85 is a pattern obtained in studying a synchronous vibrator. The actual time of one cycle is determined by finding the reciprocal of the vibrator frequency. A 100-cps vibrator requires 1/100 second or 0.01 second per cycle.



Courtesy Du Mont Labs.

Fig. 19-85.—Synchronous vibrator waveform study; TC, 100 per cent time duration of one cycle; TP_2, closure time of No. 2 primary contact; TS_2, closure time of No. 2 secondary contact; TS_1, closure time of No. 1 secondary contact; TP_1, closure time of No. 1 primary contact; V_1, vertical line caused by No. 1 contact spacing; V_2, vertical line caused by No. 2 contact spacing; O's, vertical openings caused by high velocity of voltage rise. The oscilloscope is unable to write or trace at that velocity with the given intensity control setting: O_1, vertical opening caused by No. 1 contact spacing; O_2, vertical opening caused by No. 2 contact spacing; M, "make" side of half-cycle, the "hash" corner is caused by voltage peaks due to oscillation in buffer transformer and to contact chatter; B, "break" contact side of half-cycle, indicates the secondary to primary differential peak; TD_1, differential in closure time between contact No. 1 and its respective secondary contact No. 3; TD_2, differential in closure time between contact No. 2 and its respective secondary contact No. 4.



Courtesy Du Mont Labs.

Fig. 19-86.—Various defects that may occur result in definite waveform patterns such as these, which can be interpreted to decide the probable cause of a specific trouble.

The openings in the vertical lines in Fig. 19-85 are caused by the exceptionally rapid rise in voltage that occurs immediately after the vibrator contacts meet. In this portion of the cycle, the voltage change is many times faster than in other portions of the cycle. Therefore, a light trace for this portion, hardly visible, is obtained.

The various defects that may occur result in definite waveform patterns which can be interpreted to decide the probable cause of a specific trouble. Referring to Fig. 19-86, sketch 14 shows the result of an excessive differential between primary and secondary contact spacing. One or both of the secondary contact spacings are much wider than their respective primary contact spacings, causing too much lag. The secondary contacts of a synchronous vibrator should always be spaced slightly wider than the respective primary contacts, but only slightly. If too much lag exists, it will appear on the waveform as is indicated in Fig. 19-86, sketch 14. This results in much lowered efficiency of the vibrator. The differential peak which reliably indicates this condition is located at the right-hand end of the horizontal lines of the waveform. This is the "break" contact point of the cycle. Normal differential peaks are shown in the sketches of Fig. 19-84 and also at points *B* in Fig. 19-85.

In Fig. 19-86, sketch 15, insufficient differential between the primary and respective secondary contact spacing is shown. This condition is objectionable in that it causes arcing, sparking, and burning of contacts. It causes the primary contacts to open the circuit while yet carrying heavy current. When an adequate differential is obtained, the secondary contacts remove the electrical load just before the opening of the primary contacts. Also, the secondary contacts do not apply the electrical load until after the primary contacts have closed on the opposite half-cycle.

Insufficient differential or no lag at all also causes bad timing, erratic waveform, and transient voltage peaks. An example of insufficient differential is shown on the No. *1* half of Fig. 19-86, sketch 15. The No. *2* half of the same sketch shows no lag or reverse differential where actually the No. *4* secondary contacts are leading the No. *2* primary contacts.

In Fig. 19-86, sketch 16, defective contacting, as indicated by irregularities on the horizontal portion of the waveform, is shown; the cause is probably oxidation of the contact material, or possibly improper machining of the contacts so that irregular surfaces or poorly matched surfaces are obtained. The surfaces may not be making contact evenly over a sizable portion of the available area, but only at a burned point or a segment of the rim. This defect also is sometimes caused by abnormal chatter or bouncing of side contacts.

The irregularities illustrated on the horizontal lines of sketch 16 are definitely an indication of existing abnormal high resistance at power contacts. In sketch 17, excessive unbalance, one vertical opening longer than the other, is illustrated. This defect of adjustment means that one side or half of the vibrator is carrying more than its share of the load or that the current is not evenly divided on each half-cycle. The contacts may be spaced statically equal, to the eye, while not in operation, and yet be considerably unbalanced dynamically when in operation. (This condition exists when the amplitude of reed swing is faulty or provides greater duration of contacting time on one side than on the opposite side.) Correct adjustment of the contact spacing makes allowance for this condition and produces equal contact closure time on each side. This condition is illustrated in sketch 17, Fig. 19-86.

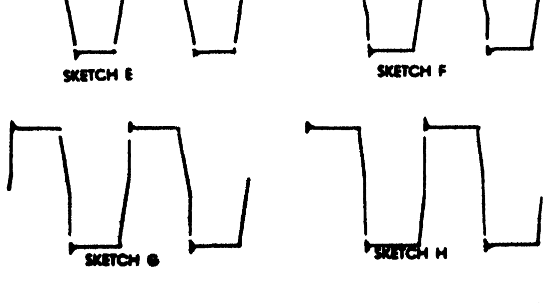
In Fig. 19-86, sketch 18, a bad retrace with excessive peaks condition is illustrated. The upper, or No. *1* contact, half of the waveform shows excessive peaks caused by insufficient differential, imperfect matching of contact surfaces, or too close spacing. The lower, or No. *2* contact, half shows a bad retrace, sometimes caused by an erratic, unsteady swing or reed, or a varying amplitude of reed swing. This is sometimes caused by dirty or pitted driving-coil contacts, an intermittently shorted driving coil, or excessive chatter of side contact plates.

Nonsynchronous Vibrator

The preceding discussion dealt with the synchronous vibrator. In general, the same principles of operation and analysis apply to the nonsynchronous vibrator which is operated in conjunction with a rectifier tube and filter system. However, the nonsynchronous type has no secondary contacts to adjust and has, therefore, no contact differential or lag. This fact indicates the waveform changes only slightly at the ends of the horizontal lines. The differential peaks ("dog ears") which are prominent on a synchronous waveform, as illustrated in Fig. 19-84, are missing in Fig. 19-87. This illustration shows four different adjustments of a nonsynchronous vibrator.

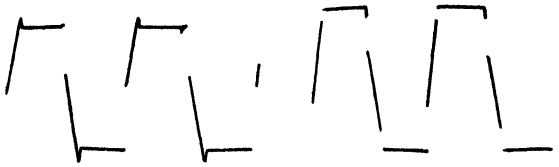
Fig. 19-87, sketch E, illustrates the effect of comparatively wide contact spacing — approximately 36 per cent contact closure time. This spacing is used on vibrators having high-voltage or high-power applications. (Note the very small openings in the vertical lines.)

Fig. 19-87F illustrates the waveform of contacts adjusted somewhat closer than those in Fig. 19-87E, approximately 40 per cent closure time. (Note that higher contact closure time of this adjustment produces less slope to vertical lines than that shown in Fig. 19-87E.) Fig. 19-87G illustrates closely spaced contacts to give a closure time of 43 per cent. Fig. 19-87H



Courtesy Du Mont Labs.

Fig. 19-87.—Nonsynchronous vibrator waveforms under different contact spacings and closure times.



Courtesy Du Mont Labs.

Fig. 19-88.—Oscillograms of normal operation of the two vibrator types, placed side by side for comparison.

shows the effect of very close spacing, with 46 per cent closure. Note that the horizontal lines are longer, and vertical lines are nearly straight up and down. Fig. 19-88 shows oscillograms of normal operation of the two vibrator types, placed side by side for comparison.

Checking Signal Generators

The cathode-ray oscilloscope is very useful for checking the waveforms of signals from the lowest audio frequencies up to radio frequencies as high as 10 Mc. In general, however, its principal use is for checking waveforms in the audio range and at moderately high radio frequencies.

The use of the oscilloscope in television work will be covered in the television section of this book. The other practical uses of the instrument will be covered in this section. The usual application of the oscilloscope in audio work involves connecting the instrument at the *output* of an amplifier system, as shown in Fig. 19-89. The dummy load resistance is usually very small in comparison with the input impedance of the oscilloscope, so that loading of the circuit under test is negligible. To check distortion, it is assumed that the waveform of the output signal provided by the generator is a sine wave or a square wave, as the case may be. Actually, this assumption can not always be made with safety.

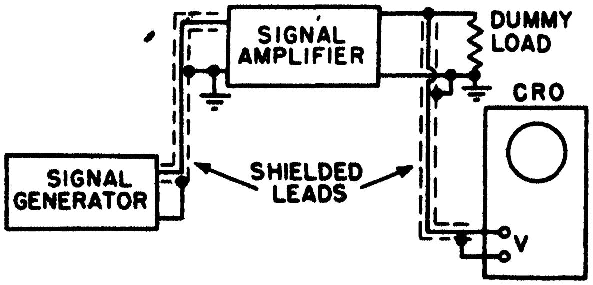


Fig. 19-89.—Test setup for observation of signal generator output.

While modern generator design has improved to the point where a sine-wave output is the standard, it is good practice to check the generator by connecting the oscilloscope across its output, preferably with the generator connected to the amplifier input. The waveform is then noted. If the waveform at the output is different, distortion has been introduced by the intervening amplifier.

As the frequency of the generator is made higher and higher, it becomes an r-f generator. Usually, the average commercial oscilloscope will have an upper frequency limit no higher than 0.1 Mc to 1.0 Mc. Some laboratory-type oscilloscopes may have a top limit as high as 5 to 10 Mc, so that the utility of the oscilloscope, as an indicating instrument, is usually limited to low frequencies in the audio range and to a-m broadcast radio frequencies.

Radio-Frequency Tests

The usual procedure in checking an r-f generator is to employ a special wideband amplifier or laboratory oscilloscope. Few servicemen, however, have the desire or need to check apparatus in that way; they depend, instead, upon the manufacturer of the test equipment to do the job and design the signal generator properly. Therefore, in this discussion, we are going to limit ourselves to audio work. The output of an f-m or television generator can be checked easily, if the serviceman is interested, by feeding the signal into a reasonably good f-m or video detector in the conventional manner. Such generators, however, if new, may usually be depended upon to provide proper waveforms, particularly when they bear the name of a reliable manufacturer. When there is some doubt about the generator's characteristics, or if the equipment possibly has been damaged and it is found desirable to check it, a standard f-m or television set in good condition may be used with the generator for waveform analysis.

F-m and television alignment is covered in a separate section of this book and information on these subjects is also given elsewhere in Rider publications.[2,3] However, in the course of our investigations, some appropriate material of interest has been developed and will be shown here.

Fig. 19-89 shows the basic circuit, whether a.f. or r.f. is involved. Usually, the r-f generator output is one volt or less and may be measured in microvolts or milli-

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[2] Rider, J. F., and Uslan, S. D., "FM Transmission and Reception," John R. Rider Publisher, Inc., New York, N. Y., 1948.

[3] Rider, J. F., "Television How It Works," John F. Rider Publisher, Inc., New York, N. Y., 1948.

volts. If its output is too high, it may seriously overload the receiver. Yet, if the output is low, no indication will be obtained on the oscilloscope. Therefore, the oscilloscope is not very useful as an r-f signal tracer and test instrument unless a special r-f probe is used in conjunction with it (except in transmitter circuits when the r-f gain required is zero and direct connections to the vertical deflection can be made).

In the audio range, the signal voltage may also be high, except at the input to a high-gain audio amplifier or at a low-level voltage amplifier stage in such an amplifier. In checking equipment where the voltages are low and the circuit impedances are high, it is better, as a general principle, to use loose coupling between the generator stage by stage, leaving the oscilloscope connected to the output circuit of the equipment, than to move the oscilloscope input about the equipment circuits in the manner of a signal tracer.

In checking an r-f generator, the test circuit of Fig. 19-89 may be used. The r-f output, at 1,000 kc or lower, is fed into the vertical amplifier input of the oscilloscope. Short, direct connections are best. It may be necessary to turn the vertical gain control of the oscilloscope all the way up and to use the maximum output of the generator to get a trace on the screen. If the generator is the standard a-m type modulated at 30 per cent with a 400-cps audio note, the sweep of the oscilloscope may be set at 200 cps for a two-cycle pattern. Internal synchronizing may be used. However, a portion of the modulating voltage may be picked off with a suitable potentiometer, connected through a 0.1-μf capacitor to the modulator, and be fed to the external "sync" input of the oscilloscope. Using a laboratory-type generator with a variable percentage of audio modulation, the crests and valleys in the modulated wave envelope can easily be seen as the technician adjusts the percentage of modulation. In some experimental tests, a number of patterns were obtained. Fig. 19-90A shows the pattern resulting from the application of a 100-kc unmodulated r-f carrier to the circuit. Note the streaks in the pattern. Each of these streaks indicates a kink in the waveform of the signal and denotes that the wave is not pure and contains harmonics. This is shown in Fig. 19-90B where several cycles of this 100-kc carrier voltage are spread by the sweep in the oscilloscope. Note the departure of this carrier voltage waveform from the ideal sine wave.

Fig. 19-90C shows the pattern for an unmodulated r-f carrier of 100 kc, but which is quite free of harmonics. Note the absence of streaks in the pattern. A few cycles of this voltage are shown in Fig. 19-90D. This method of judging the character of an unmodulated carrier, within its limitations, is fast and accurate, but not necessarily quantitative with respect to the exact amount of distortion present in the voltage wave.

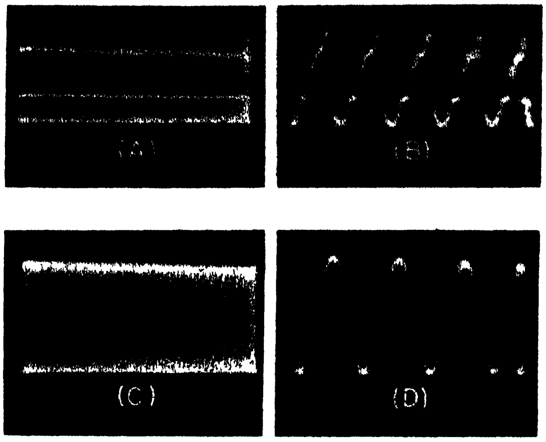


Fig. 19-90.—(A) Block pattern of nonsinusoidal wave; (B) shows a few cycles of (A); (C) block pattern of sine wave; (D) shows a few cycles of (C).

In connection with modulated wave patterns, it should be realized that there is a tendency for the waveform to change with frequency, since the percentage of modulation cannot be held constant. However, in modern well-designed generators, the effect is minimized.

Another factor to bear in mind is that the waveform of the modulating signal of the generator may not be identical with the waveform of the output voltage of the second detector in a receiver. The reasons are that a certain amount of nonlinearity and distortion may be present in both the modulator and demodulator circuits.

In Fig. 19-91A, the wave envelope of the modulated output of a good commercial signal generator, such as

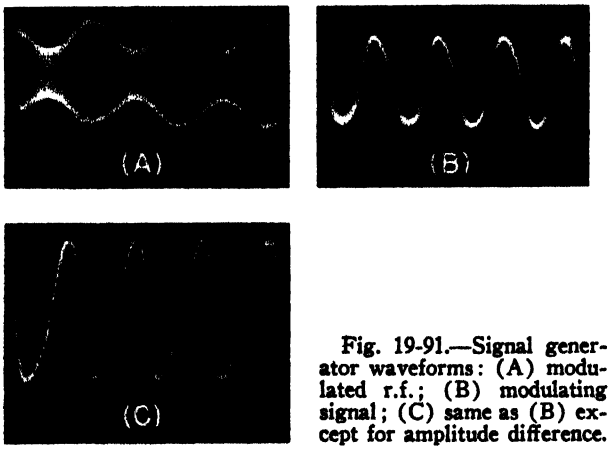


Fig. 19-91.—Signal generator waveforms: (A) modulated r.f.; (B) modulating signal; (C) same as (B) except for amplitude difference.

that used by servicemen, is shown. The audio output of the instrument, illustrated in Fig. 19-91B and Fig. 19-91C, shows the audio demodulated signal. The audio waveforms are reasonably alike, differing principally in amplitude. Since taking these oscillograms, further developments in generator and receiver design have made it difficult to discern any great difference in the audio patterns. Fig. 19-92A shows the wave envelope of the modulated output of a commercial signal generator; Fig. 19-92B, the a-f output of the instrument, and Fig. 19-92C the demodulated (audio) signal. The signal generator, in this case, is either improperly adjusted or of poor quality.

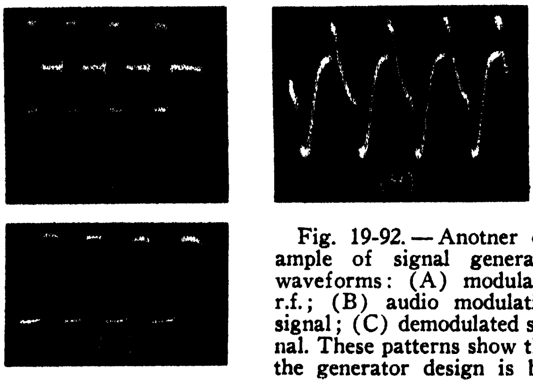


Fig. 19-92. — Anotner example of signal generator waveforms: (A) modulated r.f.; (B) audio modulating signal; (C) demodulated signal. These patterns show that the generator design is bad or that the instrument is malfunctioning due to a defect.

Distortion resulting from the use of a poorly designed or defective generator is illustrated in Fig. 19-93. At (A), the wave envelope of the modulated output of the generator is shown. At (B), the audio output of the instrument is indicated, and at (C), the demodulated waveform is depicted. It is badly distorted, containing many harmonics, and approaches a square wave in character. A poor receiver may give similar distortion, even if the signal generator is working perfectly.

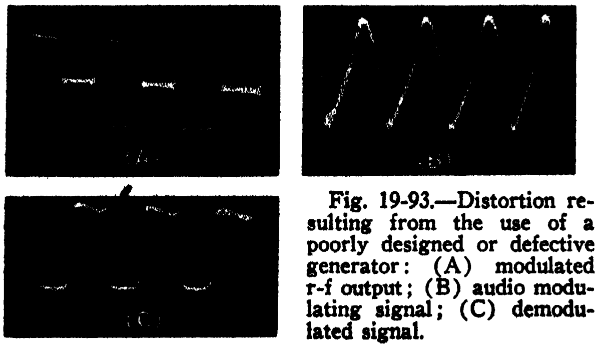


Fig. 19-93.—Distortion resulting from the use of a poorly designed or defective generator: (A) modulated r-f output; (B) audio modulating signal; (C) demodulated signal.

Oscillograms of another commercial oscillator are shown in Fig. 19-94. The audio wave is shown at (A), the 260-kc modulated wave at (B), and the demodulated wave at (C).

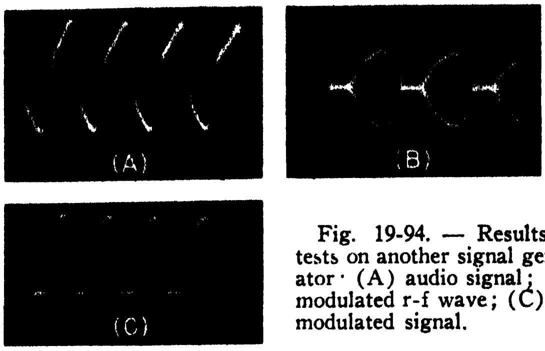


Fig. 19-94. — Results of tests on another signal generator · (A) audio signal; (B) modulated r-f wave; (C) demodulated signal.

Fig. 19-95 shows the output of a signal generator at 600 kc. Changing the oscillator frequency results in a change in waveform. Fortunately, more modern instruments, developed since the taking of the oscillograms, have superior waveform characteristics.

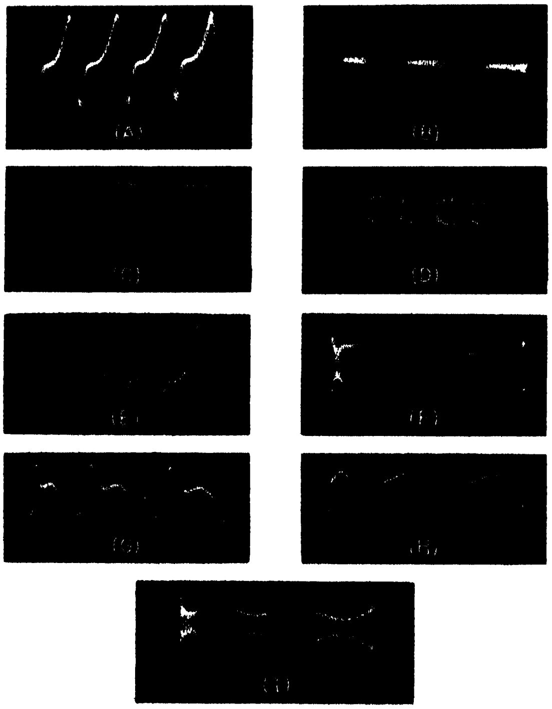


Fig. 19-95.—Waveforms of an inferior quality or poorly adjusted signal generator, different frequencies: (A) audio modulating signal for all; (B), (C) modulated and demodulated wave, respectively, for a carrier of 260 kc; (D), (E) same as (B), (C), except frequency is 600 kc; (F), (G) same as (B), (C), but 1,000 kc; (H), (I), same as (B), (C) but 1,700 kc.

Some of the old style generators still in-use can be modified for improved waveform by a radio expert having the necessary technical background. Modern instruments are well designed and should not be tampered with. Fig. 19-96 shows, at (A), (B), and (C), three oscillograms obtained from an audio oscillator in which regeneration was excessive and the L/C ratio

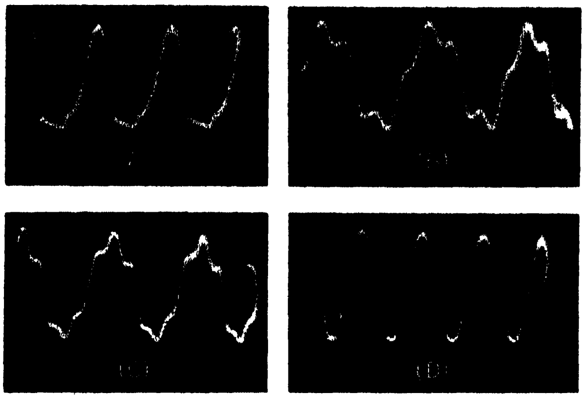


Fig. 19-96.—Audio oscillator oscillograms: (A), (B), (C), show effects of excessive regeneration and high L/C ratio. (D) shows improvement when regeneration and L/C ratio are reduced to proper values.

was very high. Reduction of the L/C ratio and regeneration resulted in the development of the sine wave in Fig. 19-96D.

An added example of the application of the oscilloscope as a tool which may be used for circuit improvement is illustrated by Fig. 19-97. At (A) is shown the pattern upon the screen when a distorted 90-kc unmodulated carrier is observed, with the linear sweep set to a very low frequency so that a block pattern results. It is possible to discern distortion by observation of the pattern. If it is solid, without any of the streaks of greater intensity, such as appear in Fig. 19-97A, the waveform usually is a sine wave or a close approach to it. The presence of streaks indicates distortion. The wave shown in Fig. 19-97A, when synchronized with the linear sweep of the oscilloscope so that a few cycles of the carrier appear on the screen, is shown in Fig. 19-97B.

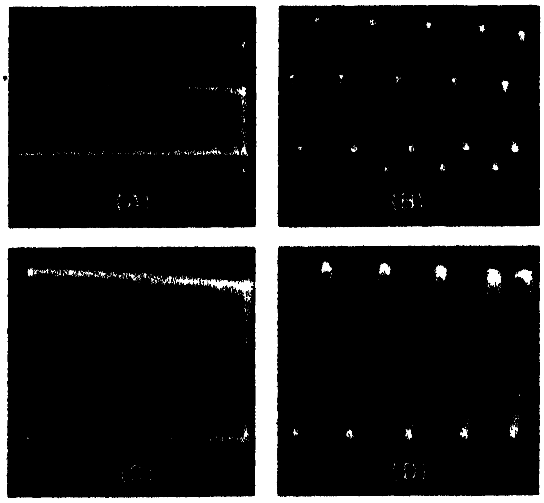


Fig. 19-97.—(A) R-f block pattern produced by using a slow sweep. (B) Complex wave used for vertical deflection of (A). Note horizontal bands resulting from secondary peaks of the wave. (C), (D) Corresponding to (A) and (B), but a sinusoidal vertical deflection is used. Note the advance of horizontal bands in (C).

Corrective remedies applied to the oscillator in the form of a load placed upon the tickler winding, so that the amount of feedback in the circuit is reduced, are shown at (C) and (D). (C) shows that the distortion has been removed from the carrier, because the streaks have disappeared. (D) shows the corresponding carrier wave, which has little of the distortion exhibited at (B).

Later on, in succeeding chapters, audio- and radio-frequency tests, using various types of generators and oscilloscopic methods, will be discussed; but at this time we may find it expedient, as well as interesting, to examine some of the industrial applications of the cathode-ray tube and the oscilloscope.

ENGINEERING AND SCIENTIFIC APPLICATIONS OF THE CATHODE-RAY OSCILLOSCOPE

The cathode-ray oscilloscope lends itself admirably to a variety of industrial and scientific applications.

Hydraulic Tests

Fundamentally, the cathode-ray oscilloscope is a voltage-actuated device. To interpret variations in fluid pressure, in oil or water, for example, a special device to change fluid pressure into electrical voltage is necessary. This device is called a "pickup." The basic relationship is shown in Fig. 19-98. The fluid pressure variations, which may be periodic or transient, slow or rapid, are converted into electrical pressures (voltages) that can be applied to the vertical input circuit of the oscilloscope.

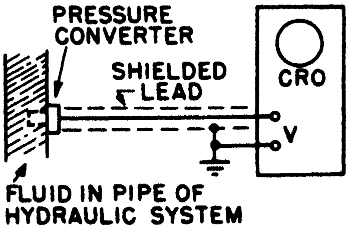


Fig. 19-98.—Use of oscilloscope in hydraulic tests.

One form of pickup, used by the Aeroquip Corporation, depends upon a variable resistance principle. This variable is one arm of a bridge circuit, normally at balance, and connected in the input circuit of the oscilloscope. An unbalance due to a resistance change causes the development of a potential which can be magnified by the vertical amplifier of the oscilloscope to give a suitable indicating pattern on the face of the tube. The bridge is in the input circuit of the oscilloscope.[4]

The equipment is somewhat analogous to vibration apparatus in which a vibration pickup is used to convert mechanical vibrations into electrical voltages. The basic vibration test idea is shown in Fig. 19-99. The pickup may use a piezoelectric crystal or an electromechanical arrangement similar to a dynamic microphone. The dynamic type is usually favored because of its comparative ruggedness.

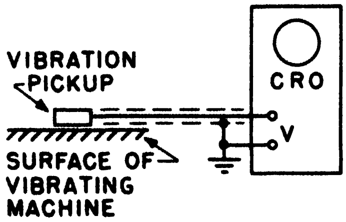


Fig. 19-99.—Setup for vibration tests with the oscilloscope.

Engine Pressure Analysis

Vibration pickups of the magnetostriction type are used in engine analysis work. Fig. 19-100 shows a basic setup. The engine vibration pickup is inserted in the cylinder and develops an electrical voltage which is fed to the oscilloscope input circuit. A pictorial block diagram for apparatus made by the Sperry Gyroscope Company is shown in Fig. 19-101. It consists of an engine, junction box, three-phase generator, amplifier, indicator, condition selector switch, and cylinder cycle selector switch. A functional block diagram is shown in Fig. 19-102. The simplified circuit of the synchronizing voltage generator is given in Fig. 10-103.[5] The following information is from the Sperry Gyroscope Co.

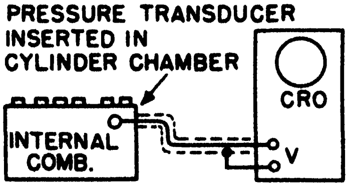
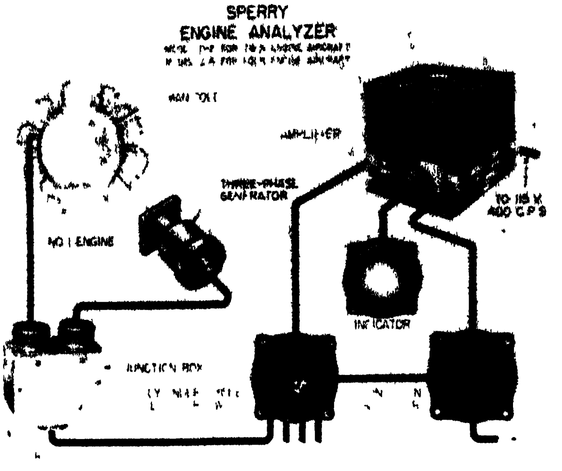
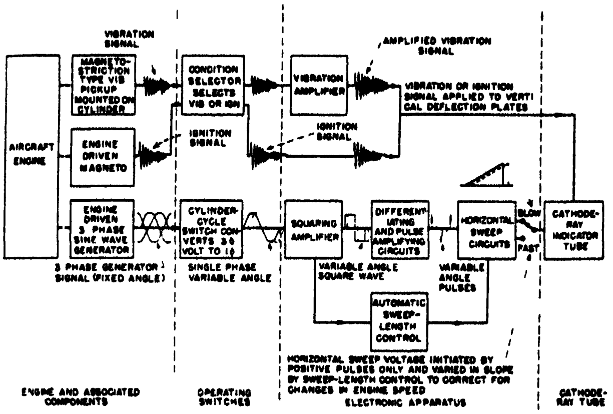


Fig. 19-100. — Engine analysis with the oscilloscope.



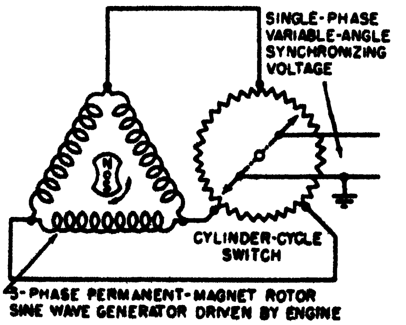
Courtesy Sperry Gyroscope Co.

Fig. 19-101.—Pictorial block diagram of the Sperry engine analyzer.



Courtesy Electronics

Fig. 19-102.—Functional block diagram, Sperry engine analyzer.


SINGLE-PHASE
VARIABLE-ANGLE
SYNCHRONIZING
VOLTAGE
CYLINDER-CYCLE
SWITCH
3-PHASE PERMANENT-MAGNET ROTOR
SINE WAVE GENERATOR DRIVEN BY ENGINE

Courtesy Electronics

Fig. 19-103.—Simplified circuit of the synchronizing voltage generator.

The two principal functions of the Sperry apparatus are the location and the identification of ignition faults and vibration defects. The oscilloscope or indicator unit of the engine analyzer is the equivalent of a special cathode-ray oscilloscope. It is designed to display pictures of voltage waveforms, to show selected details of engine operation. The voltages to be displayed are taken from the primary circuit of the ignition system, the vibration pickups on the engine cylinder, or other voltage detectors, depending on the engine functions to be observed.

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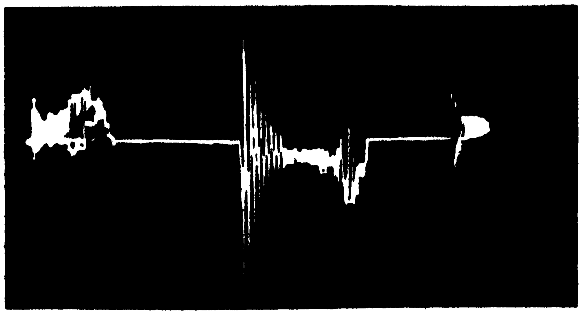
[4] An analysis of the Aeroquip oscilloscope is given elsewhere in this book. This instrument was specially designed for hydraulic measurements.

[5] Petrone, C. V., "Airborne engine analyzer," *Electronics*, vol. 21, p. 90, March, 1948.

Ignition Patterns

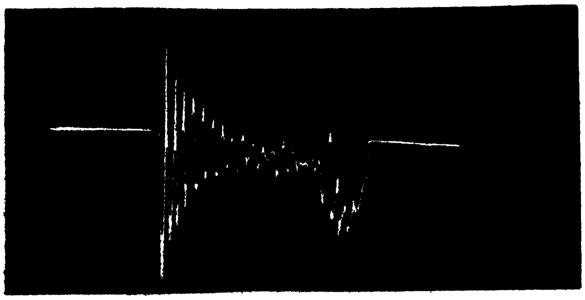
A reference pattern is obtained for normal cylinder operation. A departure from it is an indication of engine malfunctioning. Particular patterns indicate definite faults, permitting a speeding up of engine maintenance.

The ignition patterns shown here are presented by courtesy of the Sperry Gyroscope Company, and this technical information is based on material supplied by that company. The engine used for the tests was a Wright Cyclone 18, with no secondary capacitor in the magneto circuit. Although low-tension ignition systems and systems containing a capacitor in the secondary circuit produce patterns different from those shown here, the patterns are generally indicative of certain troubles. By running tests on other types of systems, which are known to be in normal working order, or to have definite defects, suitable reference patterns may be taken and photographed. Thus, there are no narrow limitations on the applicability of the electronic engine analyzer.



Courtesy Sperry Gyroscope Co.

Fig. 19-104.—Normal ignition oscillogram, Sperry engine analyzer, (engine running, 1,700 rpm).



Courtesy Sperry Gyroscope Co.

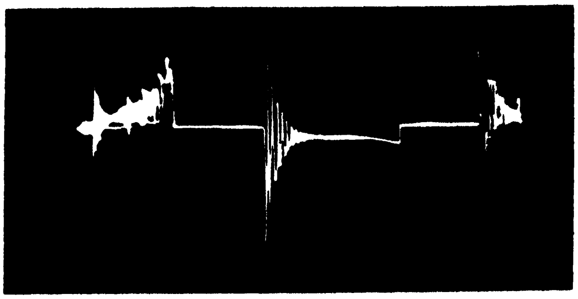
Fig. 19-105.—Low-frequency ignition pattern.

The *normal pattern* is limited in length by the opening and closing of the magneto breaker points. A straight line or zero line, which connects consecutive patterns, is obtained during the period when the points are closed. This pattern is illustrated in Fig. 19-104, taken with an engine speed of 1,700 revolutions per minute.

The *low-frequency pattern* is shown in Fig. 19-105. Because the voltage of the magneto alternates in polarity on consecutive firings, consecutive cylinders produce patterns alternately above and below a zero reference line.

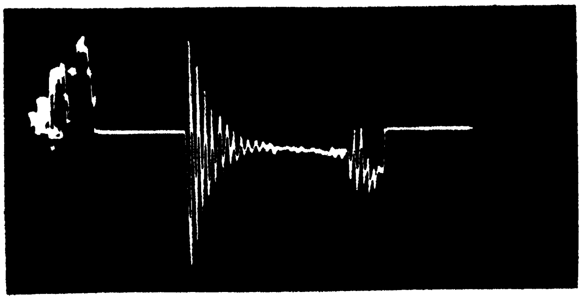
The *shorted secondary pattern* appears in Fig. 19-106. Any short circuit or arc between the high-tension lead of the spark plug and ground produces a pattern easily distinguished from the normal.

A *small gap pattern* is shown in Fig. 19-107. The short-circuit condition is produced by a plug gap smaller than normal.



Courtesy Sperry Gyroscope Co

Fig. 19-106.—Shorted secondary pattern.

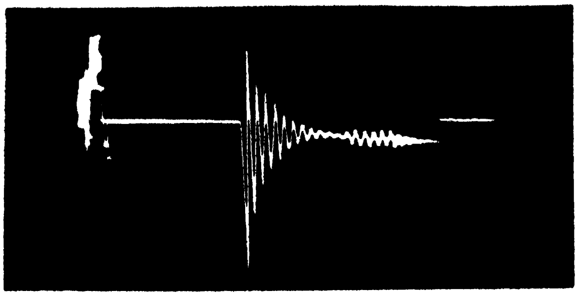


Courtesy Sperry Gyroscope Co.

Fig. 19-107.—Small gap pattern.

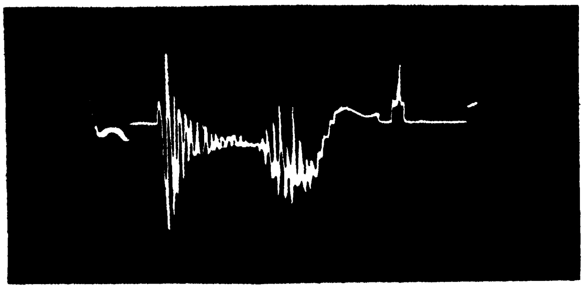
The effect of a *fouled plug* is illustrated in Fig. 19-108.

The effect of *breaker point bounce* is shown in Fig. 19-109. A bouncing action of the breaker points, on closing contact, will produce a variety of pips which make their appearance after the normal finish of the ignition pattern. The number of pips corresponds to the number of times the points bounce on closing.



Courtesy Sperry Gyroscope Co

Fig. 19-108.—Fouled plug pattern.



Courtesy Sperry Gyroscope Co

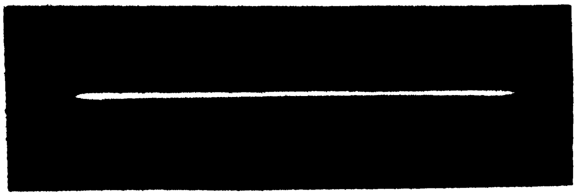
Fig. 19-109.—Effect of breaker point bounce.

The condition of a *shorted primary capacitor* results in a straight line trace, as shown in Fig. 19-110.

A certain amount of noise is to be expected. This is shown by the *engine noise pattern* of Fig. 19-111.

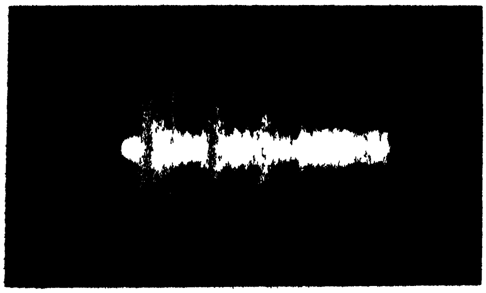
The effect of an *improperly seated intake valve* is shown in Fig. 19-112.

The condition of *no combustion* is illustrated in Figs. 19-113 and 19-114.



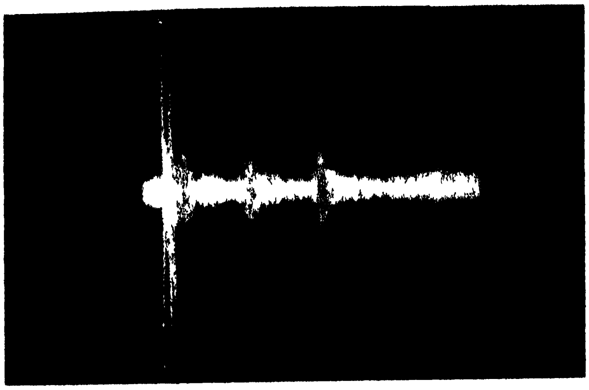
Courtesy Sperry Gyroscope Co

Fig. 19-110.—Effect of shorted primary capacitor.



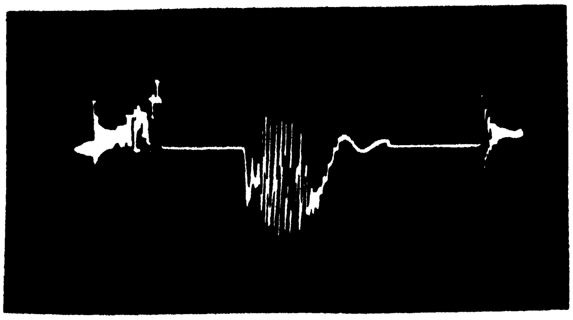
Courtesy Sperry Gyroscope Co.

Fig. 19-111.—Engine noise pattern.



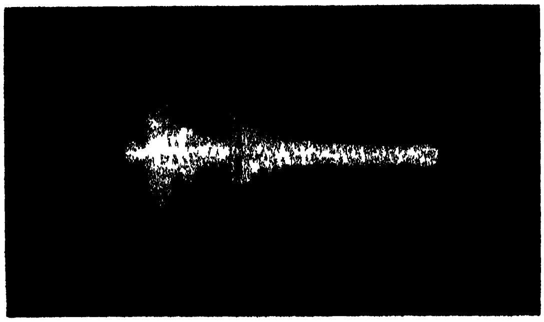
Courtesy Sperry Gyroscope Co

Fig 19-112—Effect of an improperly seated intake valve.



Courtesy Sperry Gyroscope Co

Fig. 19-113.—Lack of combustion

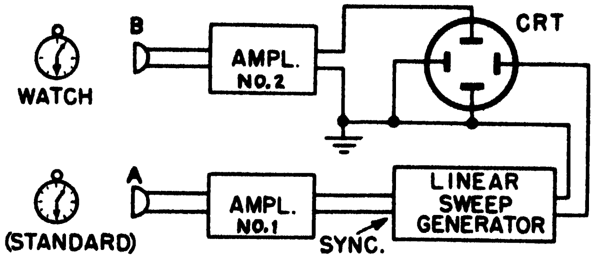


Courtesy Sperry Gyroscope Co.

Fig 19-114.—Another pattern indicative of lack of combustion.

Watch Timing

The timing of a watch can be checked by comparison with a standard by use of a circuit arrangement such as that shown in Fig. 19-115. Microphone *A* picks up sound impulses from the standard and feeds them, as electrical impulses, through audio amplifier No. *1*. The output of the amplifier synchronizes the sawtooth sweep voltage applied to the horizontal-deflection plates. Any small difference in timing will be indicated by drift of the pattern on the screen.



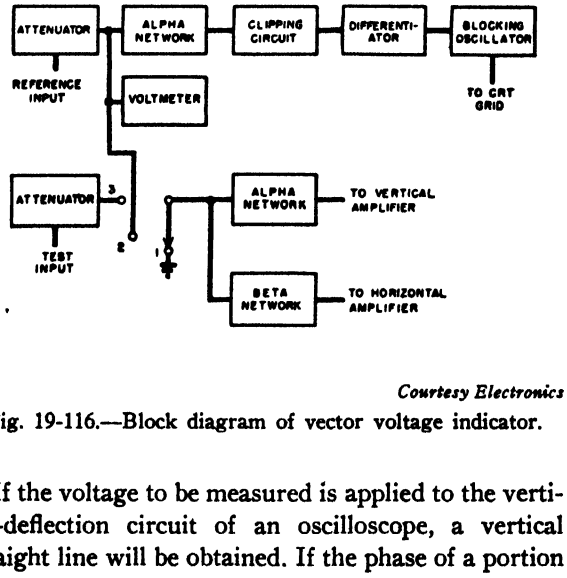
Courtesy RCA

Fig. 19-115.—Circuit for use of oscilloscope in watch timing.

Vector Voltage Indicator[6]

This material is based on a magazine article by Sulzer. Using a cathode-ray oscilloscope, with auxiliary equipment, it is feasible to show a complex quantity on the screen of a cathode-ray tube. Readings may be taken in polar or rectangular co-ordinates.

Fig. 19-116 is a block diagram of the system. The frequency range is 50 to 5,000 cps. The full-scale sensitivity is variable from 1 volt to 500 volts rms, with a phase accuracy of ±3° and an amplitude accuracy of ±10 per cent of full scale.



Courtesy Electronics

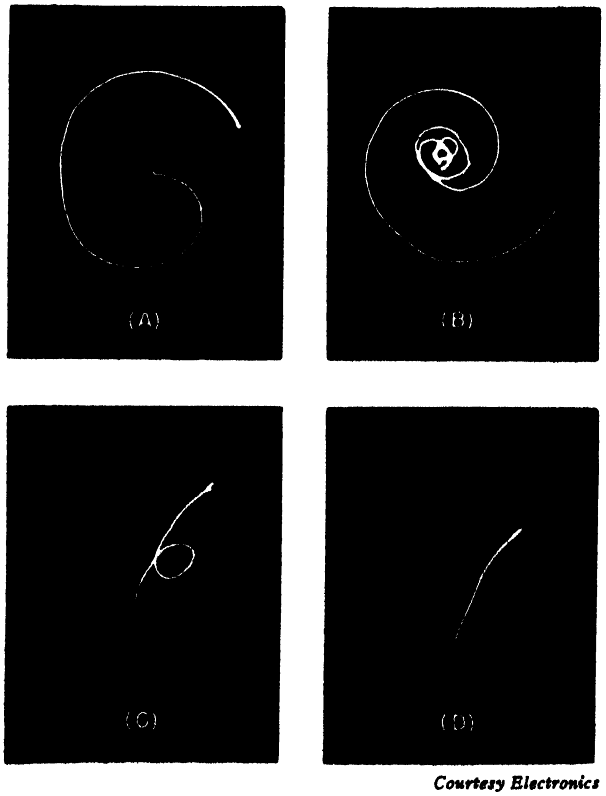
Fig. 19-116.—Block diagram of vector voltage indicator.

If the voltage to be measured is applied to the vertical-deflection circuit of an oscilloscope, a vertical straight line will be obtained. If the phase of a portion of that voltage is shifted 90° and applied to the horizontal-deflection circuit, a circular trace will be obtained with the proper gain adjustment.

It is apparent that the diameter of the circle will be proportional to the amplitude of the test voltage and can be used to indicate its magnitude. It is also apparent that the velocity of the spot tracing the circle will be constant, with the result that a linear phase scale will be obtained around the circle.

The position of zero phase angle on the circle must be established with the aid of some other voltage, here called the reference voltage. In practice, the reference voltage might be the input to a circuit under test, while the output voltage of the circuit would constitute the test voltage. The reference voltage is clipped and differentiated so that a pulse is obtained at its zero-degree position. With the circle blanked by means of the intensity control, the pulse is used to brighten the trace for a very short time. Therefore, the pattern consists of a single dot which denotes the tip of a vector drawn from the origin, which is placed at the center of the screen. If a polar co-ordinate scale is drawn on the screen, the length of this vector indicates the magnitude of the test voltage, while its angular position from the origin is the phase angle of the test voltage with respect to the reference voltage. Two phase-shifting networks are used, known as the alpha and beta networks, to produce the 90° phase shift required.

Four patterns obtained with the vector voltage indicator are shown in Fig. 19-117. The curves illustrate the action of the electron beam dot on the cathode-ray-tube screen as it is swept through its locus by changing one of the circuit parameters.



Courtesy Electronics

Fig. 19-117.—Patterns obtained with vector voltage indicator.

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[6] Sulzer, P. G., "Vector voltage indicator," *Electronics*, vol. 22, pp. 107-109, June, 1949.

Transmission-Line Measurements[7]

The study of transmission-line phenomena is facilitated by the use of the cathode-ray oscilloscope. It enables the user to see electrical transmission-line conditions and, in contrast with previous methods, allows permanent records to be made by photographic means.

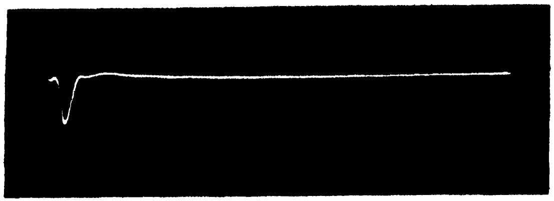
The oscilloscope features which apply specifically to transmission-line studies are the internal trigger pulse generator, driven sweep, and beam modulation amplifier. The internal trigger pulse generator furnishes an output signal of either positive or negative polarity, consisting of one-half microsecond pulses of approximately 100 volts peak amplitude, the repetition rate being variable between 200 and 3,000 pulses per second. This pulse serves as an indicator of the transmission-line properties.

A clear picture of transmission-line characteristics is made possible by using an oscilloscope having a self-contained, driven sweep. The sweep operates so that repeated pulses fall on top of each other, leaving a section of the sweep clear for the presentation of reflections.

The beam modulation amplifier provides a means of impressing the time demarcations on the pattern for quantitative or calibration purposes. This amplifier is used in conjunction with an external signal or with the internal timing oscillator. The signal is applied through the amplifier to the grid of the modulating electrode of the cathode-ray tube, and can be made to blank out the beam, resulting in accurate markers along the trace under observation. The timing oscillator provides marker signals at intervals of 1, 10, or 100 microseconds, and is keyed by the driven sweep circuit.

Method Used

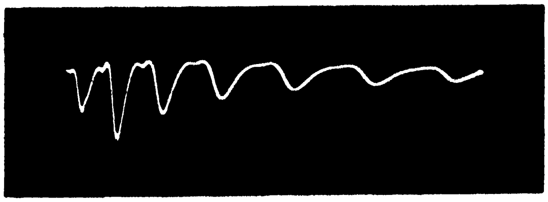
To utilize the oscilloscope in transmission-line studies, the negative trigger output of a Du Mont Type 248 oscilloscope, or equivalent, is used to synchronize the oscilloscope's driven sweep and also to supply the pulse to be observed. The negative trigger output is used so that a large pulse will merely cut off the first amplifier stage, and not drive the grid positive. As the pulse is fed to the transmission line at the vertical amplifier input terminal, and there is no external sending-end termination, the Y-axis amplifier input impedance is the sending-end terminating impedance. R_L, the receiving-end terminating impedance, is selected so that the characteristic impedance of the transmission line is within its range. Fig. 19-118 is a photograph of the pattern produced when the receiving-end impedance is equal to the characteristic impedance. It is apparent that only the original pulse is visible, since no reflections have taken place.



Courtesy Du Mont Labs.

Fig. 19-118.—Test pattern obtained in transmission-line measurements when the line is matched.

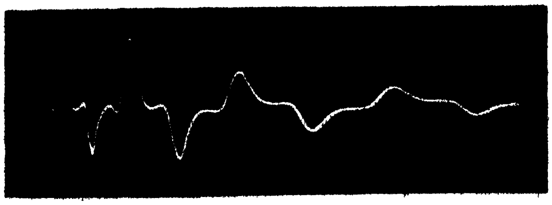
In order to determine the unknown characteristic impedance Z_0 of any transmission line, it is merely necessary to connect the line and vary the receiving-end impedance until a pattern identical to Fig. 19-118 is obtained. This adjustment is critical, since any slight variation from the characteristic impedance will produce a reflected pulse which will be visible on the oscilloscope screen.



Courtesy Du Mont Labs.

Fig. 19-119.—Pattern resulting when the terminating impedance is not equal to the characteristic impedance and R_L is greater than Z_0.

Figs. 19-119 and 19-120 illustrate patterns which result when the terminating impedance is not equal to the characteristic impedance. In the case of Fig. 19-119, R_L is greater than Z_0. In Fig. 19-120, Z_0 is greater than R_L. In both cases, the terminating impedance at the sending end is the oscilloscope Y-axis amplifier input impedance, in parallel with the pulse generator output impedance.



Courtesy Du Mont Labs.

Fig. 19-120.—Pattern resulting when the terminating impedance is not equal to the characteristic impedance and Z_0 is greater than R_L.

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[7] Ossmann, E. A., "Transmission line measurements," *Du Mont Oscillographer,* vol. 7, pp. 1-4, May-June, 1945.

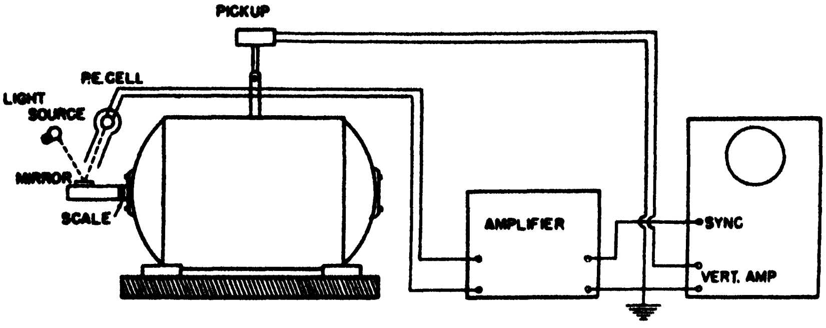

Fig. 19-121.—Dynamic balancing setup.

Courtesy Du Mont Labs

The characteristic impedance of the line is approximately 50 ohms, making the sending-end impedance much larger than the characteristic impedance. It is noted that the second pulse on both photographs is almost twice the amplitude of the first. This is a result of the terminating impedance at the sending end being much greater than the characteristic impedance.

With the patterns shown in Figs. 19-119 and 19-120, most of the important properties of transmission lines can be determined readily. Utilizing the beam modulation feature of the oscilloscope (Du Mont 248 or equivalent), we can impress a time scale on the pattern. The transmission time along the line may be determined by measuring the distance between the centers of the first and second pulses. This measurement is made by counting the time markers between the two points.

Once the transmission time is known, the velocity of propagation can be calculated since

$$V = \frac{L}{T}$$

where V is the velocity of propagation, L is the length of the line, and T is the transmission time along the line (one way).

In addition to determining the presence of a discontinuity on a transmission line, we can now locate it. Any such discontinuity along a line will create a reflection which will appear as a foreign pulse on the oscilloscope. The transmission time to the discontinuity may be determined as explained previously; and so, with the velocity of propagation known, we can solve for L, the distance to the discontinuity. We can also forecast the type of fault by observing the polarity of the reflection.

The oscilloscope is also very useful in the continuous monitoring of transmission lines to detect and locate faults immediately after their occurrence. A photograph of the normal pulse conditions on a particular line may be taken and any new discontinuity can readily be detected by placing the negative of the photograph over the oscilloscope screen.

Dynamic Balancing[8]

Vibration in a rotating machine is frequently due, in large measure, to unbalance of the rotating part. Various equipment has been designed to analyze this vibration for the purpose of locating the unbalance in order to correct it. The cathode-ray oscilloscope may be used in smaller shops where the correct balancing of high-speed motors and other machines is desired.

Two accessories are necessary. The first of these is a directional vibration pickup. Also, an arrangement is required to synchronize the oscilloscope sweep circuit with some known point on the rotating part. Perhaps the simplest scheme is a mechanical contact driven by the machine and so connected with a dry cell as to supply an impulse to the synchronizing terminal on the oscilloscope. This method may be used successfully but a photoelectric system is better, using a flicker-free light source, a narrow mirror taped to the moving part, a photoelectric cell, and an amplifier. This equipment is shown in Fig. 19-121.

With the machine under test and in motion, the oscilloscope sweep circuit is adjusted to the correct frequency and is under the control of the synchronizing circuit. The oscilloscope pattern represents one full revolution of the rotating parts. If the vibration pickup is now mounted on the frame of the machine and connected to the vertical amplifier input circuit, the resultant pattern shows the relationship between the vibration at the point of mounting, in the direction of the axis of the pickup, and the position of the rotating part. This information will indicate the position of an overweight in the moving part.

While the use of the present equipment has been used in the balancing of electric motors, the procedure is equally applicable to other rotating machinery. The convenience and reliability of the method is increased if the drive is reversible.

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[8] Grant, Wylie, "The oscillograph used for dynamic balancing," *Du Mont Oscillographer*, vol. 5, pp. 1-4, December, 1941-January, 1942. The following material is based upon this article.

Vibration

Vibration in a machine may be caused by: a simple unbalance in the rotating part; dynamic unbalance in the rotating part; extraneous causes, such as neighboring machinery, gear or belt disturbances, etc.; internal causes, such as bearing trouble, wind forces, magnetic forces, etc.

Dynamic Unbalance

An exaggerated example of dynamic unbalance is shown in Fig. 19-122. The shaft carries two weights, each of which is mounted off balance, but they balance each other when at rest. However, when the shaft rotates, the centrifugal forces of the two weights tend to give the shaft a weaving motion. Fig. 19-123 shows two diagrams of a rotor in which a dynamic unbalance exists. The diagram represents conditions of motion when the unbalance couple is effective and shows two positions of a rotor differing by half a turn. As the figures indicate, the rotor has a tendency to rotary vibration in any plane which contains the axis of the shaft. Consequently, in testing for dynamic unbalance, the pickup is mounted with its axis perpendicular to a line through the center of gravity of the moving part and in a plane also containing the axis of the shaft.

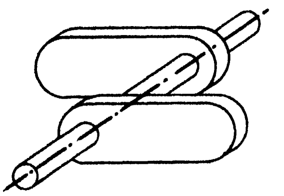
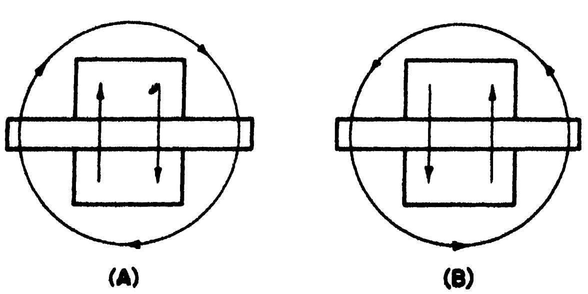


Fig. 19-122.—Exaggerated example of dynamic unbalance.

Courtesy Du Mont Labs.

Thus, the only difference in the detection of simple unbalance and of dynamic unbalance is a difference in the mounting of the pickup, which may be obtained by pivoting the pickup through 90° at a suitable bolted joint. It will be noted that the pickup, when mounted to detect one type of unbalance, will not be sensitive to the other.



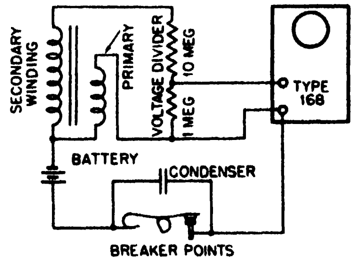
Courtesy Du Mont Labs.

Fig. 19-123.—Two diagrams of a rotor in which a dynamic unbalance exists.

With the equipment in operation, the screen pattern will probably show a single pronounced wave in the general form of a sine wave, but distorted by various higher frequencies. The single wave is the only one to be used and judgment must be used to visualize its simplified form.

Production Testing of Ignition Coils[9]

For production testing of ignition coils, the cathode-ray oscilloscope offers many advantages over earlier methods. Fig. 19-124 shows the oscilloscope arrangement for checking the output voltage of the coil in factory production testing. There is no actual spark discharge produced by the secondary. The output of the coil is readily clipped into the circuit, and the polarity and the peak voltage are quickly read from a calibrated scale, mounted over the face of the cathode-ray tube.



Courtesy Du Mont Labs.

Fig. 19-124.—Oscilloscope arrangement for checking the output voltage of the coil in factory production testing.

Fig. 19-125 shows a typical, damped secondary waveform obtained from an ignition coil by this method. If a spark gap is connected in parallel with the coil, the oscilloscope may be employed to indicate the voltage necessary to jump the gap. Also, when no safety gap is provided, this same connection may be used to serve as a maximum insulation test for the coil.

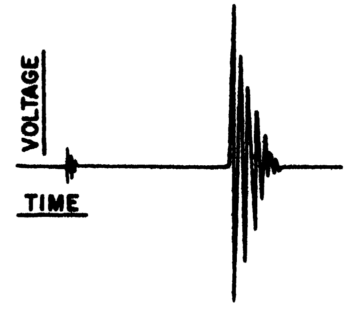


Fig. 19-125.—A typical, damped secondary waveform obtained from an ignition coil.

Courtesy Du Mont Labs.

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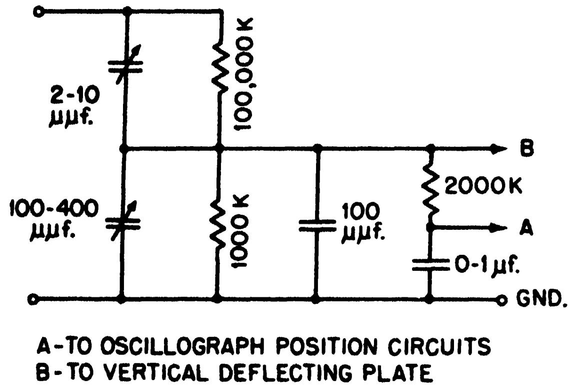
[9] Toben, G., "Production test of ignition coils with the cathode-ray oscillograph," *Du Mont Oscillographer*, vol. 3, pp. 1-4, June-July, 1939.

When an accurate waveshape plot is desired, it is necessary to use a resistance-capacitance compensated voltage divider and to deflect the cathode-ray tube directly as shown in Fig. 19-126. This type of voltage divider eliminates the errors due to stray circuit capacitances and it attenuates the high-frequency components of the signal in equal proportion for indication on the cathode-ray tube.

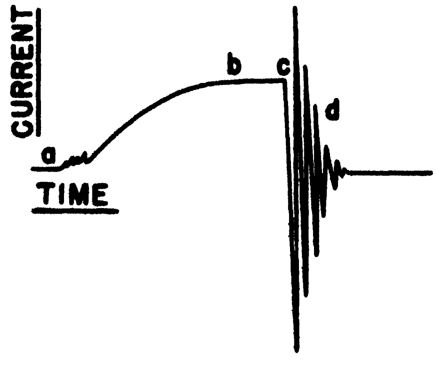

2-10 μμf.
100,000 K
B
2000K
100-400 μμf.
1000 K
100 μμf.
A
0-1 μf.
GND.
A-TO OSCILLOGRAPH POSITION CIRCUITS
B-TO VERTICAL DEFLECTING PLATE

Courtesy Du Mont Labs.

Fig. 19-126. — Resistance-capacitance coupled voltage-divider circuit.

In Fig. 19-127, the primary current may be seen to increase logarithmically from the "make" at point *a* to the saturation point *b*. The break occurs at point *c* and the oscillatory discharge is indicated from *c* to *d*.

By adjustment of the horizontal gain control, which determines the amplitude of the sweep-circuit deflection, the baseline may be adjusted until it covers exactly 36 divisions on the calibrated scale over the face of the tube. This method of operation then permits direct readings of cam angle, 10° per division, indicating the time during which the breaker points are closed, the time of saturation, and the timing of the internal combustion engine.

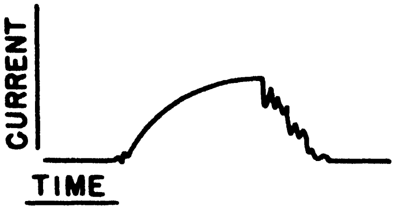


Courtesy Du Mont Labs.

Fig. 19-127.—Graph of primary current.

Although the sweep-frequency control markings do not amount to a frequency calibration of this circuit, they will give a fairly accurate indication of the engine speed. Weak spring tension or resilience in the contact support arms is indicated by a flutter when closing, as indicated at point *a*, Fig. 19-127.

At low engine speeds, the primary current will reach saturation, as shown at *b* in Fig. 19-127. At higher engine speeds, the break may occur previous to saturation and poor ignition will result. The oscillatory discharge shown from *c* to *d* in Fig. 19-127 assumes the form shown in Fig. 19-128 when there is a defective capacitor connected across the breaker points.



Courtesy Du Mont Labs.

Fig. 19-128.—Graph of oscillatory discharge when there is a defective capacitor connected across the breaker points.

Engineering Tests of Ignition Coils[10]

The preceding paragraphs have covered the use of the oscilloscope in the production testing of ignition coils. The instrument may be used to advantage in the determination of coil characteristics.

In many studies of ignition coils, it is desirable to determine the turns ratio of a given coil. This may be conveniently done with the oscilloscope, using the connections shown in Fig. 19-129. The resistance divider, connected across the secondary of the ignition coil in this figure, has a voltage division very closely equal to 2,000:1. An input potential of approximately 1 volt, rms, at power-supply frequency, is applied to the primary of the ignition coil. The switch is thrown to position *B* and the deflection, on the oscilloscope, will then correspond to 1/2,000 of the output voltage. The gain control of the oscilloscope should be adjusted for a convenient signal deflection *Y1* about ⅝ of full screen, and the total deflection should be measured on the calibrated scale of the instrument. The switch should then be thrown to position *A*, and the total deflection should be recorded without disturbing the setting of the oscilloscope gain control.

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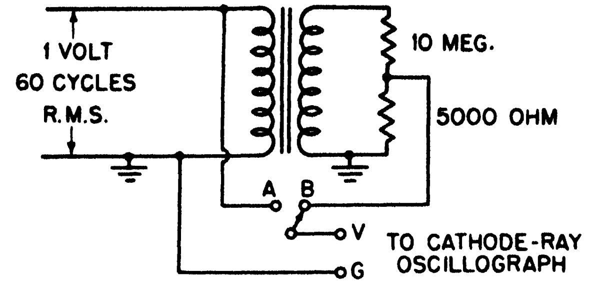
[10] "Oscillograph-study of ignition-coil characteristics," *Du Mont Oscillographer*, vols. 3 and 1, pp. 1-4, February-March, 1940.



Courtesy Du Mont Labs.

Fig. 19-129.—Circuit for determination of turns ratio of ignition-coil transformer.

When *N1* indicates the number of primary turns, *N2* the number of secondary turns, and the turns ratio is designated by *N1/N2*, the turns ratio may be calculated from

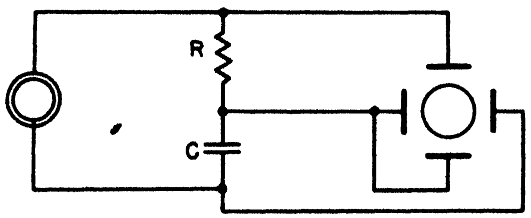
$$\frac{N1}{N2}=\frac{Y1}{2{,}000\,Y2}.$$

The figure 2,000 is obtained from the attenuation ratio of the voltage divider, and should be changed accordingly for different attenuation ratios.

Cathode-Ray-Oscilloscope Stopwatch[11]

The oscilloscope, with no auxiliary equipment other than an ordinary still camera and a few inexpensive circuit components, may be used readily to measure time intervals of short duration with an accuracy well under 0.1 millisecond.

In the design, adjustment, or testing of signaling systems, it is frequently essential to be able to measure the speed with which certain relays, switches, or other devices operate. Most relays and snap switches, excepting those embodying some special sort of retard feature, function within the range of about 2 to 15 milliseconds.



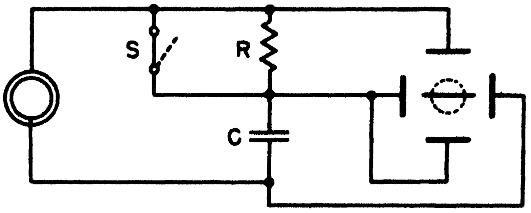
Courtesy Du Mont Labs.

Fig. 19-130.—Phase-splitting circuit for development of circular pattern.

Method

The method to be described is based upon the principle that if two sine voltages of identical frequency and amplitude but having a phase difference of 90° are applied, one to the vertical and one to the horizontal deflection plates of a cathode-ray tube, the resultant spot motion will be a theoretically perfect circle, will have a uniform rate of rotation, and will make one complete revolution for each cycle of the deflection voltages. Such a circular pattern may be developed by the phase-splitting circuit shown in Fig. 19-130. The resistance *R* must equal the reactance of *C* at the frequency employed.

Suppose we now insert a switch *S*, as shown in Fig. 19-131, into the circuit. So long as the switch *S* is open, both the deflection voltages are applied to the deflection plates, and the spot travels its circular path. When switch *S* is closed, the voltage drop across the resistor is eliminated, so that there is no vertical deflection, and the spot oscillates to and fro, in a horizontal path, under the driving force of the voltage across the capacitor *C*.



Courtesy Du Mont Labs.

Fig. 19-131.—Introduction of switch to preceding circuit results in this circuit.

Transit-Time Measurements

Suppose now it is desired to measure the transit time of a relay. The relay contacts are connected in place of the switch *S* as shown in Fig. 19-132. A still camera, which may be of any convenient type, having a reasonably fast lens, is focused upon the screen of the cathode-ray tube. With the relay at rest, a straight line appears on the screen. When the relay operates and its contacts separate, the spot jumps from this "base" line and starts to travel in a circle. When the relay transit is completed and the contacts again close, the spot jumps back to the baseline. The pattern, as thus photographed, will consist of a straight line and an arc, which may or may not meet the straight line, depending upon the time in the cycle when the operation occurred.

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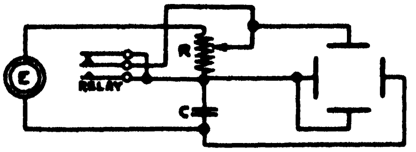
[11] Brailsford, H. D., "CRO stopwatch," *Du Mont Oscillographer*, vol. 4, pp. 1-4, June-July, 1940.



FOR E=60~ R=50,000 C=0.1μf

Courtesy Du Mont Labs.

Fig. 19-132.—Relay contacts connected in place of switch *S* of Fig. 19-131.

In making the exposure, care should be taken to open the shutter immediately before operation of the relay and to close it immediately after. If this is not done, an over-exposure is likely to fog the film, due to the brilliance of the horizontal line, which is on the screen continuously except for the interval during which the contacts of the relay are actually separated.

The described method is not limited to the testing of relays and switches, but may be applied also to the investigation of moving machinery parts if the construction is such that the auxiliary electrical contacts can be fitted. The circular pattern may also be used to check the timing of a camera shutter. A short-persistence screen is necessary for this purpose, to prevent the introduction of errors in measurement due to the "afterglow" of the screen material, and the procedure is simply to make a snapshot of the circular pattern. The photograph will show an arc of a circle of a length corresponding to the time during which the shutter was open.

Single-Sweep Stop Watch[12]

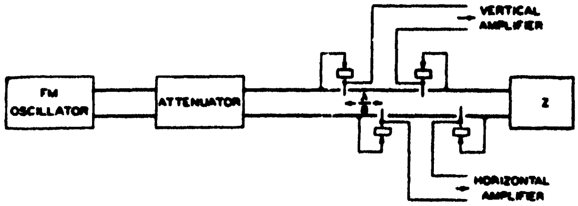
In essence, the operation of this equipment is as follows: The first of two events whose time separation is desired is used to trip the single-sweep circuit of an oscilloscope. A 60-cps signal, obtained from the power line and applied to the vertical-deflection circuit of the instrument, is used to provide a time standard. The second event is used to cause a potential difference of several volts to be impressed instantaneously on the vertical-deflection circuit, thus abruptly terminating the trace by deflecting it off the screen. The vertical length of the signal retained on the screen is then used as a measure of the time lapse between the two events.

Special Impedance Method[13]

An experimental study at ultra-high frequencies of the input impedance as a function of frequency for any given circuit element requires measurements of the complex impedance at a number of different frequencies over a prescribed frequency range. The oscilloscope offers a convenient and rapid method of presenting the data directly, in the desired form, without computation. The principal advantage of this method is one of speed. It is possible to observe the variations in the input impedance of devices under transient conditions.

Circuit

The block diagram of the simple four-probe method of obtaining test information is shown in Fig. 19-133. Four probes are used to sample the waves existing in the transmission line, which may be either a waveguide or a coaxial line. These probes are in pairs, the two probes of each pair being spaced along the line by a quarter wavelength, while the two sets are staggered by an eighth wavelength.



Courtesy Proc. I.R.E.

Fig. 19-133.—Block diagram of simple four-probe method of obtaining test information.

The output from each probe goes directly to a crystal detector. The two crystals of one pair of probes are balanced, the difference in their outputs being impressed on one pair of deflection vanes in the oscilloscope. The difference in the outputs from the second pair is impressed in a like manner on the other pair of deflection vanes in the oscilloscope. If, then, the input to the transmission line is varied in frequency but not in amplitude, and if the crystals follow a square law, the oscilloscope spot will trace a path representing the desired curve on the reflection-coefficient plane. The center of the reflection-coefficient plane is located by interrupting the output of the driving oscillator at the end of each frequency excursion.

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[12] Sandow, Alexander, "Single-sweep stop watch," *Du Mont Oscillographer*, vol. 4, pp. 1-4, December, 1940-March, 1941.

[13] Samuel, A. L., "An oscillographic method of presenting impedances on the reflection-coefficient plane," *Proc. I.R.E.*, vol. 35, pp. 1279-1283, November, 1947.

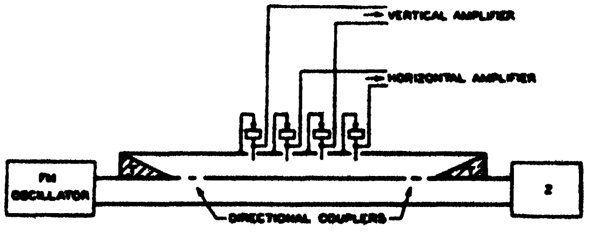
V_i is the amplitude of the incident voltage and V_r the voltage of the reflected wave. When the frequency is varied over a restricted range, the spot will trace the value of V_r in magnitude and phase. The proper scale factor for the oscilloscope and its accompanying amplifier can readily be obtained by terminating the transmission line in a short circuit, so that $V_r = V_i$, and calling the accompanying deflection unity. Since, by definition, V_r for unity V_i is the reflection coefficient, the desired relationship has been verified.

The four-probe method, however, is subject to a number of limitations and errors. Since use is made of fixed spacings in the transmission line, the method is limited in its frequency range if reasonable accuracy is to be expected. The errors will be somewhat greater for a waveguide system because of the more rapid variation in waveguide wavelength with frequency. Care must be taken to insure that the probes are not large enough to produce noticeable distortions in the standing-wave pattern which is being sampled. A third type of error is introduced if the crystals do not follow a square law in operation. This imparts a nonlinear radial scale to the oscilloscopes. This error will be largest for large values of the reflection coefficient. It will decrease as the reflection coefficient approaches zero.

With suitable attention to detail, the four-probe method has been found to yield results which are accurate to the order of 5 per cent over a 5 per cent frequency band. The crystal errors can be substantially eliminated by any arrangement which attenuates the reflected-wave component so that it is small compared with the direct-wave component.

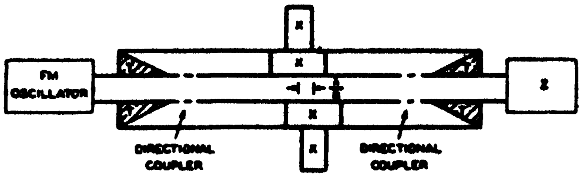
Directional Coupler

In the circuit of Fig. 19-134, two directional couplers are used to sample the incident and reflected waves. The coupler which samples the reflected wave may be constructed to have a greater coupling attenuation than the coupling attenuation of the coupler sampling the incident wave, or a resistive attenuator may be used to provide the desired difference in level. The probes in the secondary line then measure a reflection coefficient which is reduced by some desired factor with respect to the reflection coefficient of the impedance under test. Suitable allowance for this can be made in the calibration of the oscilloscope so that the device is still direct-reading. It is, of course, possible to dispense with one of the directional couplers, but the use of a single coupler to sample both wave components is not recommended because of the severe requirements on impedance matching which this usage imposes.



Courtesy Proc. I.R.E.

Fig. 19-134.—Two directional couplers are used here to sample the incident and reflected waves.



Courtesy Proc. I.R.E.

Fig. 19-135.—Waveguides, used as directional couplers, permit sampling of waves.

The use of directional couplers to sample the waves makes it possible to dispense with the use of probes and use hybrid junctions to obtain the necessary additions and subtractions. Such a circuit, using waveguides, is shown in Fig. 19-135. Two input couplers and two output couplers are shown. The output from one incident-wave coupler and one reflected-wave coupler are added and subtracted by one hybrid tee with a crystal on each of the two remaining arms. The second hybrid tee is displaced by an eighth wavelength to obtain the necessary phase difference. The balanced crystal outputs go to the oscilloscope amplifiers. The sensitivity of the directional coupler and the hybrid junction method can be greater than that of the simple probe method, since it is possible to obtain much larger samples of the incident wave, without serious error, by these means.

Elliptically Polarized Light[14]

With relatively simple auxiliary equipment, elliptically polarized light may be demonstrated on the oscilloscope screen.

A beam of elliptically polarized light is usually most easily understood when it is considered as the resultant of the combination of two beams of plane-polarized light, these components being oriented perpendicular to each other, and representing sinusoidal vibrations of the same frequency with a definite time lag or phase difference between corresponding points on the two waves.

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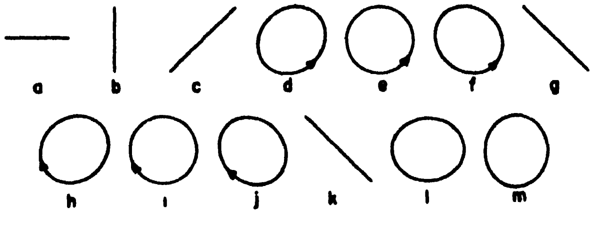
[14] Steinhardt, R. Laurence, "The cathode-ray oscillograph as a means of demonstrating elliptically polarized light," *Du Mont Oscillographer*, vol. 4, pp. 1-4, October-November, 1940.

By purely electrical means, it is possible to produce oscilloscope patterns which possess geometrical properties identical to these same properties in a beam of polarized light. For the purposes of demonstration, these patterns may be taken as true representations of certain states of polarization. For example, if a sinusoidally varying voltage, such as the 60-cps lighting source, is impressed on the horizontal plates of the oscilloscope, the "spot" will move rapidly back and forth to produce a horizontal line, the length of which is proportional to the magnitude of the impressed voltage.

Similarly, a vertical line of controlled length may be produced if the voltage is applied to the other set of deflection plates. Considering the oscilloscope screen as a cross section looking down into the beam of elliptically polarized light, these two lines may be taken as representations of the two basic, perpendicularly plane-polarized components of the ellipse *a* and *b* as seen in Fig. 19-136.



Courtesy Du Mont Labs.

Fig. 19-136.—Patterns produced in study of elliptically polarized light.

If these two components are now combined, and if they happen to be of equal magnitude and of the same phase (or "in step"), they will produce, as a resultant pattern, a line at 45° to the two components, as shown at *c* in Fig. 19-136. If the magnitude of the two components remains the same, but their phase angle difference increases, the pattern on the screen successively changes to a left-handed ellipse with the major axis at 45° *d*, then a left-handed circle when the phase difference becomes 90° at *e*.

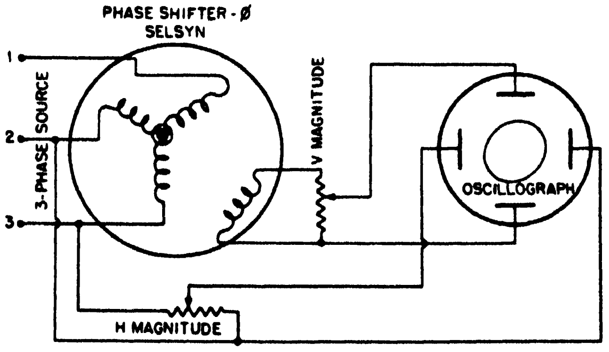
Further increasing of the phase difference results in a pattern consisting of a left-handed ellipse inclined at 135° *f* which degenerates into a straight line at 135° when this quantity approaches 180°. As the phase difference becomes greater than 180°, these patterns are then repeated in the reverse order exactly as before, except that the direction of rotation is now right-handed *h* to *k*.

If a beam of plane-polarized light enters a photoelectric specimen under mechanical stress, and if the polarization azimuth is at 45° to the lines of electrical or mechanical stress, the beam will undergo a continuous change as it penetrates deeper into the stressed material.

By altering the magnitude of one or the other of the two components with the phase difference set at 90° or 270°, it is possible to produce ellipses inclined at 0° or 90° (*i*, *l*, and *m*), and, by varying all three elements, it is possible to represent any imaginable state of polarization, and to note directly the magnitude of each component and the phase difference.

Method

All the patterns previously described can be produced with quite simple electrical equipment such as is usually found in most physics laboratories. The principal item is the cathode-ray oscilloscope, which is preferably of the electrostatic-deflection type. For this purpose, the Du Mont 164 oscilloscope, or equivalent, with a three-inch screen, is suitable.



Courtesy Du Mont Labs

Fig. 19-137.—Wiring diagram of Selsyn motor 3-phase circuit.

A phase-changing transformer, to be operated from a three-phase source, so as to produce a voltage of constant magnitude and of controllable phase, is required.

From the wiring diagram in Fig. 19-137, it will be seen that a Selsyn motor has its three-wire winding connected to a three-phase source of the proper voltage and its output winding connected, via a potentiometer, to the vertical-deflection plates of the oscilloscope. Similarly, one phase of the input is connected, via another potentiometer, to the horizontal-deflection plates.

With an oscilloscope, the built-in amplifiers and "V-gain" and "H-gain" controls will take the place of the two potentiometers.

If the source frequency is 50 or 60 cps, the pattern-forming "spot" will move around the screen too fast for the eye to perceive whether the patterns are rotating left-handed or right-handed, and it will be necessary to resort to some trick to demonstrate the transition from left-handed to right-handed elliptically polarized light when the phase angle passes through 0° or 180°.

Application

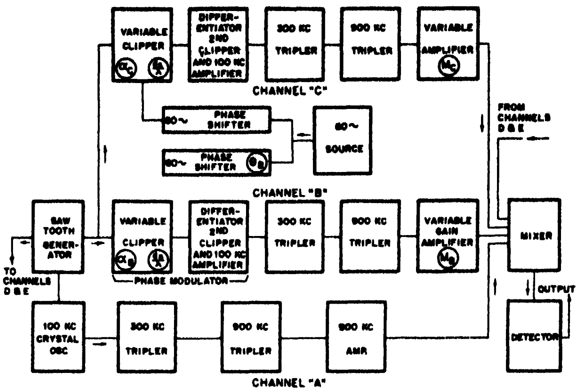
The two gain controls determine the magnitude of the two perpendicular sinusoidal (plane-polarized) components, while the Selsyn motor or phase-changing transformer alters the phase relationship of these two components without affecting the magnitude of either. By equalizing the magnitudes, it is possible to demonstrate all the patterns in Fig. 19-136*c* to 19-136*k* with one rotation of the Selsyn motor shaft, and the relationship of plane-polarized light, with both left-handed and right-handed rotation, is quite evident.

RCA Antennalyzer[15]

The equations for the radiation patterns of directional antenna systems are well-known, but the arithmetical work necessary to secure a plot of the radiation pattern is tedious and time-consuming. Using this equipment, much work in antenna design is eliminated.

In general, the designer of a directional antenna for broadcast station use knows the pattern required and is faced with the problem of determining the antenna configuration which will yield this pattern. The RCA Antennalyzer was developed to synthesize or to analyze. The instrument is entirely electrical, with no moving parts, except the potentiometers which change the various parameters. Developed specifically for the design of directional antennas for broadcast use, the Antennalyzer, as constructed, will yield the radiation pattern of directional antennas which have as many as five towers or sources of radiation.

Each source is characterized by four parameters: (1) the distance from a reference point; (2) the azimuth angle with respect to a baseline; (3) the amount of current in the antenna; (4) the phase angle of this antenna current. Thus, the Antennalyzer has four potentiometers associated with each antenna, with one exception. One antenna is located at the reference point and carries unit current at zero phase. Hence, no controls are required for this antenna.



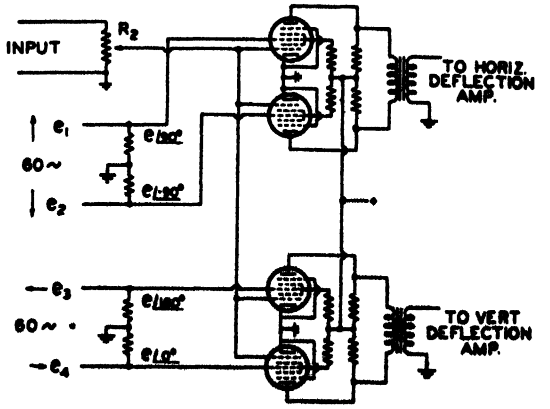
Courtesy RCA and Proc. I.R E.

Fig. 19-138.—Block diagram of RCA antennalyzer.

Pattern Display

The radiation pattern is displayed directly on the face of a cathode-ray tube, either in polar or rectangular coordinates. The Antennalyzer may be used in two ways. The dials may be set to correspond to a given antenna configuration. Then, the resulting pattern is observed on the cathode-ray tube. However, when a given pattern is the goal, the dials may be twiddled until the proper pattern is obtained. Then the dial settings are recorded. These dial settings tell where to locate the towers, as well as the current ratios and phase angles to use. With a little practice, this operation of analysis may be performed in a few minutes. Metering devices are included in the Antennalyzer so that the ratio of maximum field intensity to rms field intensity is obtained. The block diagram of the Antennalyzer is shown in Fig. 19-138.

The complete, mathematical theory underlying the operation of the equipment is given in the original article.[15]



Courtesy RCA and Proc. I.R.E.

Fig. 19-139.—Circuit for development of polar pattern.

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[15] Brown, H. G., Morrison, C. W., "The RCA antennalyzer—an instrument useful in the design of directional antenna systems," *Proc. I.R.E.*, vol. 34, pp. 992-999, December, 1946.

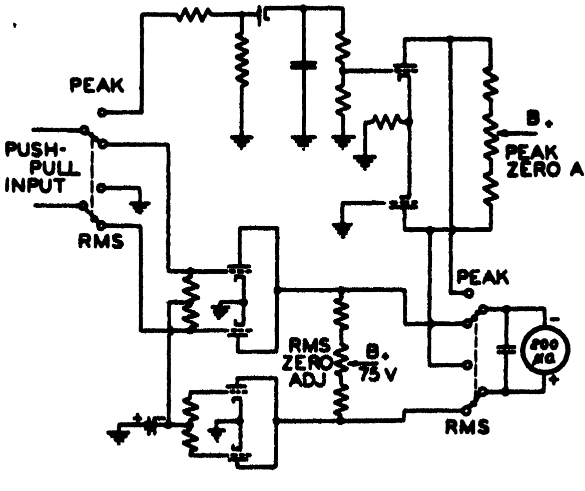
To display the pattern in rectangular co-ordinates is by far the simplest procedure. However, many engineers are used to looking at the patterns plotted in polar co-ordinates. To provide a polar pattern, the circuit of Fig. 19-139 may be used.

D.A. Gain

The gain of a D.A. (directional antenna) may be checked with the Antennalyzer and suitable auxiliary apparatus. When the antenna designer has arrived at a pattern of the proper shape, he usually desires a knowledge of the scale factor to place on the plot. An exact determination involves a knowledge of the mutual resistances existing between the antennas. However, an approximate answer may be obtained quickly by plotting the pattern in polar co-ordinates and measuring the area of the pattern with a planimeter. A circle whose area is the same is then taken as the circle from a single antenna operated with the same power. The radius of this circle is the rms value of the horizontal polar diagram. This approximate relation may be obtained by measuring the peak value of the signal coming out of the Antennalyzer and then measuring the rms value of the same signal.

The signal from the Antennalyzer is fed to the input terminals shown in the upper left-hand corner of Fig. 19-140. First, the signal is fed to the peak-reading voltmeter, which consists essentially of a diode rectifier. The input signal is adjusted until the meter in the output reads full scale. Then, the signal is switched to the input of the rms voltmeter. The output meter is switched to this voltmeter at the same time. Since the meter scale is calibrated with full scale equal to unity, the new reading is now the ratio of the rms signal to the peak signal.

The rms voltmeter is a square-law device. Since the output of the Antennalyzer contains a d-c component as well as a-c terms, cross-product terms not properly included in the determination of the rms of the pattern are brought into the reading unless the complex circuit with push-pull input and push-push output is used.



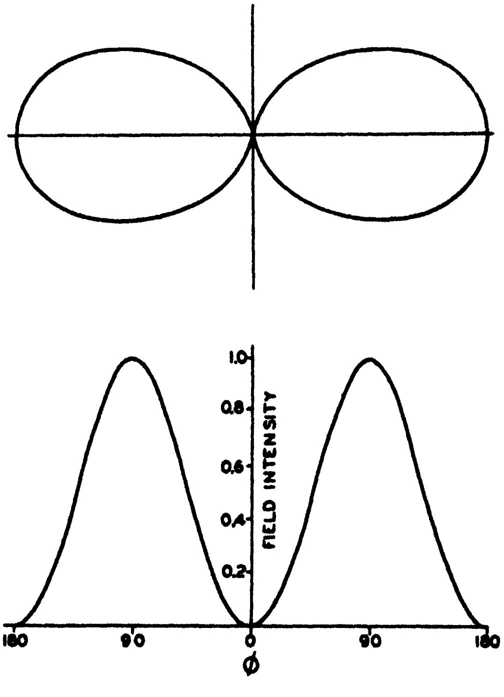
Courtesy RCA and Proc. I.R.E.

Fig. 19-140.—The signal from the antennalyzer is fed to the input terminals shown in the upper left-hand corner of this drawing.

Operation

The calibration of the instrument is readily accomplished. First, all antennas are turned off except for antenna *A*. In rectangular co-ordinates, this source then gives a straight, horizontal line, displaced from the zero axis. Then, the spacing control for antenna *B* is set at zero. This places antenna *B* so that it coincides with antenna *A*.

The current-ratio dial for antenna *B* is turned on and the current ratio and the phase angle adjusted simultaneously until the horizontal line has reached a minimum position, which should coincide with the zero axis. This means that the current in antenna *B* is exactly equal and 180° out of phase with the current in antenna *A*.



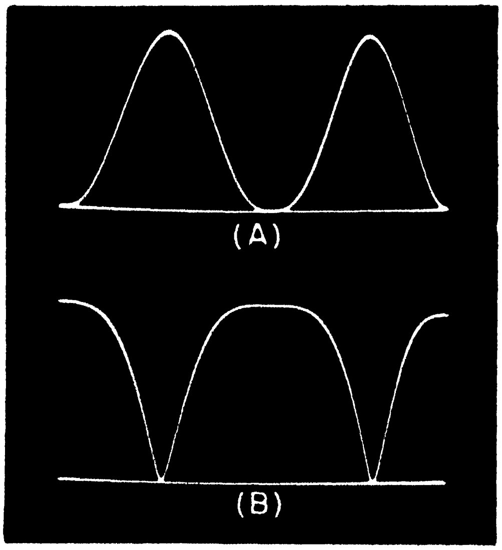
Courtesy RCA and Proc. I.R.E.

Fig. 19-141.—Calculated polar diagram for two towers carrying equal and in-phase currents, when the tower spacing is one half-wavelength.

This adjustment then establishes the unity value point on the current-ratio dial, and locates the 180° point on the phase-angle dial. The phase-angle dial is then turned until maximum displacement is found for the horizontal line. This procedure locates the zero-degree point on the phase-angle dial. Next, with antenna *B* set at unit current and zero phase, the spacing dial is turned until a "figure 8" pattern with the minimum just reaching zero is obtained. This establishes the half-wave-spacing point on the dial.

The calibration of the tower azimuth angle is easily obtained by turning this dial and suitably marking the dial as the pattern is marched across the scale. Reference to a chart of patterns[16] obtained with two towers assists in crosschecking quickly a number of calibration points.

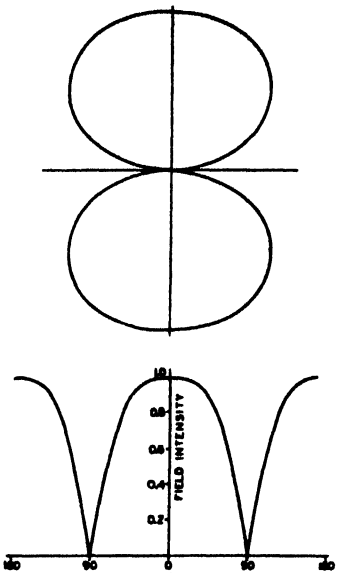
Fig. 19-141 shows a calculated polar diagram for two towers carrying equal and in-phase currents, when the tower spacing is one-half wavelength. This field pattern, plotted in rectangular co-ordinates, is also shown in the same illustration. The increase in detail of the display in rectangular co-ordinates is evident. Fig. 19-142 shows the trace on the Antennalyzer for the same set of conditions.



Courtesy RCA and Proc. I.R.E.

Fig. 19-142.—Trace on the antennalyzer for conditions of Fig. 19-141.

Fig. 19-143 shows the calculated polar and rectangular trace for the same conditions as Fig. 19-141, except that the two currents are now 180° apart in phase.



Courtesy RCA and Proc. I.R.E.

Fig. 19-143.—Calculated polar and rectangular trace for the same conditions as Fig. 19-141, except that the two currents are now 180° apart in phase.

Fig. 19-144 is the corresponding trace shown on the Antennalyzer. The polar plots of Fig. 19-141 and Fig. 19-143 fail to show the marked difference in behavior in the null points that are shown by rectangular traces.

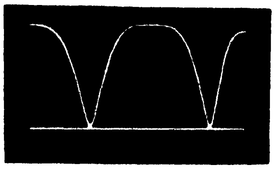
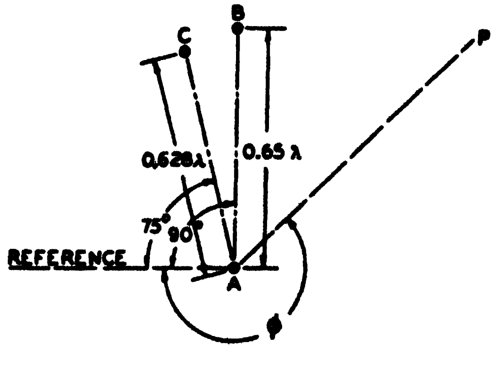


Fig. 19-144.—Corresponding antennalyzer trace for conditions of Fig. 19-143.

Courtesy RCA and Proc. I.R.E.

Fig. 19-145 shows the configuration of antennas installed at station WTAR, Norfolk, Virginia. The calculated horizontal radiation pattern is shown in Fig. 19-146, plotted in both polar and rectangular co-ordi-



Courtesy RCA and Proc. I.R.E.

Fig. 19-145.—Configuration of antennas installed at WTAR, Norfolk, Va.

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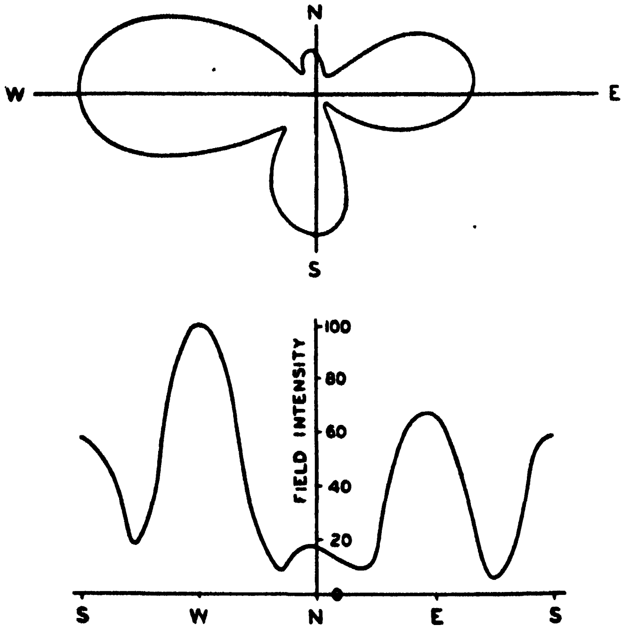
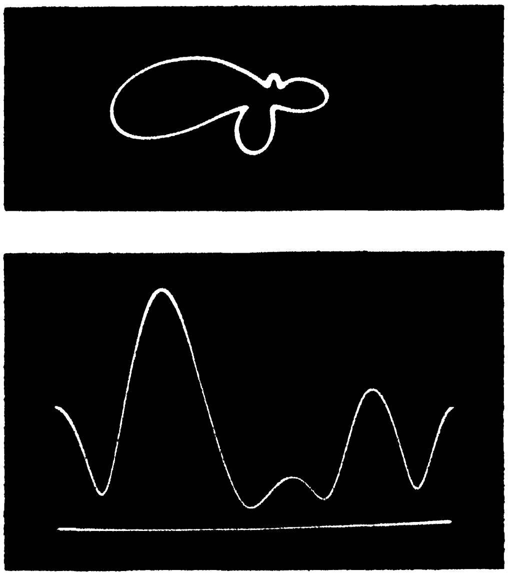
[16] Brown, G. H., "Directional antennas," *Proc. I.R.E.*, vol. 25, pp. 78-145, January, 1937.



Courtesy RCA and Proc. I.R.E.

Fig. 19-146. — Calculated horizontal radiation pattern, plotted in both polar and rectangular co-ordinates.



Courtesy RCA and Proc. I.R.E.

Fig. 19-147.—Polar plot shown in antennalyzer for conditions of Fig. 19-146.

nates. The corresponding polar plot shown in the Antennălyzer is reproduced in Fig. 19-147, while Fig. 19-148 shows the same plot in rectangular co-ordinates.

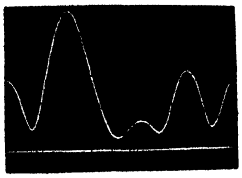


Fig. 19-148. — Same plot as in Fig. 19-147, but in rectangular co-ordinates.

Courtesy RCA and Proc. I.R.E.

Horizontal plane patterns have been shown. However, calculations and patterns in the vertical plane may be handled equally well.

Circulated Pulse Testing[17]

Square-wave and pulse-testing techniques may be extended to permit the signal pulse to be observed after circulating many times through the transmission system under test. This method is particularly useful for measuring the cumulative effect of a number of similar units, such as those used in carrier or microwave-radio-repeater systems, when only one unit is available. Applications to video-frequency, intermediate-frequency, and radio-frequency testing with a-m or f-m transient signals are feasible in checking distortion.

Basic Circuit

A block diagram of the fundamental circuit for circulated-pulse testing is shown in Fig. 19-149. The functioning of the equipment is almost self-explanatory from the drawing. The transient response of a system may be observed directly using this technique.

A simple method of observing the transient response of an amplifier or transmission system is to connect a pulse-modulated signal source to its input and an oscilloscope to its output. If the oscilloscope sweep is synchronized with the pulse rate, a stationary picture of the response to pulse signals is obtained on the screen. Since the observed response is a complicated function of what might be called the fundamental steady-state characteristics of the apparatus being tested, such methods are most valuable for comparing the different designs and for over-all qualitative tests of complete communications systems using television, facsimile, or frequency modulation.

With reference to Fig. 19-149, the transmission circuit under test is inserted in a loop consisting of a delay line, an attenuator, and an auxiliary amplifier indicated by the rectangle marked "gated amplifier." The test pulse is introduced into the loop at the input to the

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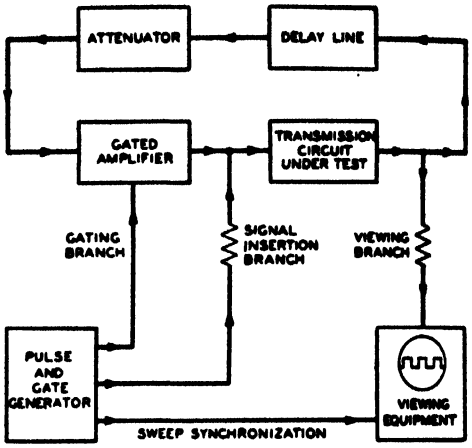
[17] Beck, A. C., Ring, D. H., "Testing repeaters with circulated pulses," *Proc. I.R.E.*, vol. 35, pp. 1226-1230, November, 1947.



Courtesy Proc. I.R.E.

Fig. 19-149.—Block diagram of fundamental circuit for circulated pulse testing.

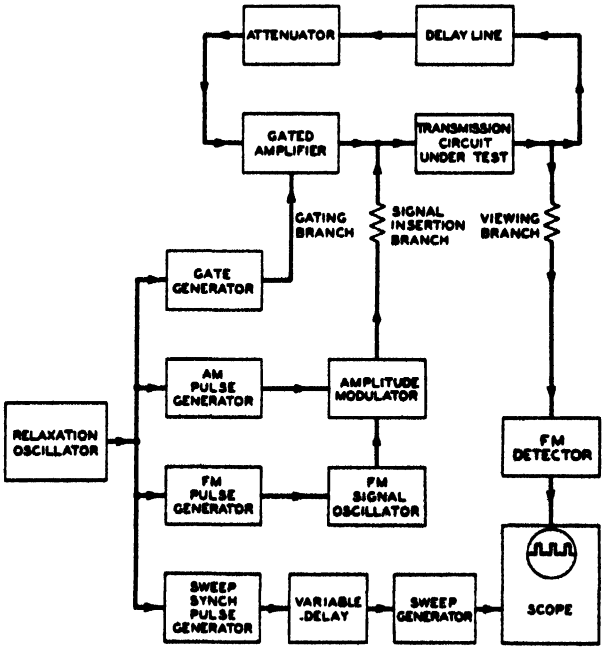
equipment under test, and a sample of the output is removed from the loop and connected to the viewing circuit. The major portion of the output is fed through the delay line, attenuator, and gated amplifier, back into the input.

The delay in the loop circuit is made greater than the duration of the test pulse, and the gain around the loop is made unity. Under these conditions, a single pulse in the loop will circulate around the loop indefinitely, and the viewing oscilloscope will show a succession of pulses which are each spaced along the sweep by the total delay time of the loop circuit. Each successive pulse on the oscilloscope shows the original pulse after another trip around the loop, and, therefore, another trip through the amplifier or system under test.

If the test pulse is made recurrent, in order to obtain a stationary picture built up of many superposed traces on the viewing oscilloscope, trouble will be encountered because pulse number one will still be circulating around the loop when pulse number two is inserted. It is the function of the gated amplifier to obviate this difficulty. This amplifier is arranged so that its gain is normally much less than unity. Just before a test pulse is inserted into the loop, a "gate" voltage is applied to the gated amplifier. This gate voltage raises the loop gain to unity and holds it there for a period which is many times greater than the loop delay, but less than the test-pulse repetition time. Thus, the test pulse continues to circulate for the duration of the gate signal. When the gate voltage is removed, the loop gain drops to much less than unity and the circulating pulse is quickly damped out so that the loop circuit is free from signals when the gate voltage is reapplied and a new test pulse is introduced.

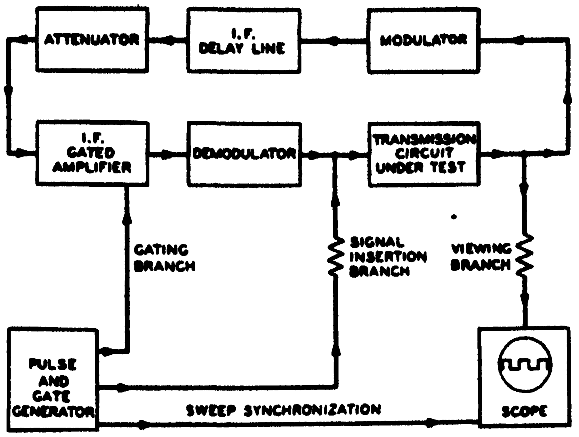
The points at which the signal insertion and viewing branches are connected to the loop are arbitrary. As shown in Fig. 19-149, the signal is inserted at a low-level point in the loop, so that a relatively small pulse voltage is required. The viewing branch is shown connected to a high-level point, so that less amplification is necessary to get a satisfactory oscilloscope deflection. When this is done, the test pulse cannot be observed until it has made one trip through the amplifier or circuit under test. If enough high-quality amplification is provided in the viewing branch, both branches can be connected to the loop at the same point, and one can monitor the test-pulse shape directly. On the block diagram, resistors are shown in the test branches to indicate that the connections are made by bridging circuits which do not disturb the impedance terminations of the equipment in the loop.

Circulated-pulse testing with frequency-modulated pulses has been carried out, as shown in the block diagram of Fig. 19-150. A block diagram of the arrangement used for circulated-pulse testing with amplitude modulation at r.f. and i.f. is similar to Fig. 19-150, except as indicated in the caption.



Courtesy Proc. I.R.E.

Fig. 19-150.—Block diagram of circuit used for circulated pulse testing, using f-m pulses. For a-m testing the f-m pulse generator is removed, and the signal oscillator and the detector are changed to a-m types.



Courtesy Proc. I.R.E.

Fig. 19-151.—Circuit arrangement for video frequency testing.

Video-Frequency Testing

A wide-band video circulating loop requires a high degree of balance in the gated amplifier, in order to avoid transients which circulate with the test pulses. It also requires a precision delay line which has uniform attenuation and delay over the necessary video band. The various problems may be avoided using the arrangement shown in Fig. 19-151. which was devised by W. M. Goodall of Bell Laboratories. A frequency modulator and discriminator are connected to the video system to be tested, with the gating and delay accomplished in the i-f part of the loop. Amplitude modulation and detection could be used in the same way.

Memory Oscilloscope[18]

The memory oscilloscope is used for obtaining photographic records of randomly occurring electric transients. This instrument differs from other oscilloscopes of the memory type in that its operation is completely automatic. Forty sets of oscillograms can be recorded without the aid of an operator. Because of this feature, such an oscilloscope may be used for monitoring electrical systems in which sporadic circuit disturbances are to be studied. The unit consists essentially of three cathode-ray tubes with associated power supply, a sawtooth oscillator to provide the horizontal time axis, amplifiers for the vertical-deflection circuits, a beam-blanking circuit, and an automatic camera.

When the memory oscilloscope is in operation, the cathode-ray tubes are continuously excited. The vertical-deflection circuits of the tubes are connected to the voltages to be investigated, and each tube screen carries a cathode-ray pattern characteristic of the voltage applied to its vertical-deflection plates. Upon the occurrence of a transient pulse, the camera shutter is released and the electronic beams of the cathode-ray tubes are interrupted. The camera then records the persistence of fluorescence, which is characteristic of the material used on the tube screens.

In this way, a record is obtained of electric events that occurred previous to the opening of the shutter. To insure successful operation, the component parts of the automatic memory oscilloscope must be electrically co-ordinated and controlled.

Control Circuit

Fig. 19-152 is a block diagram illustrating the sequence of functions performed by the automatic memory oscilloscope control circuit. The control is initiated by a voltage pulse applied to the grid of an electron tube. This pulse may be supplied by the transient voltage itself—or some other voltage pulse which occurs simultaneously with the transient to be studied.

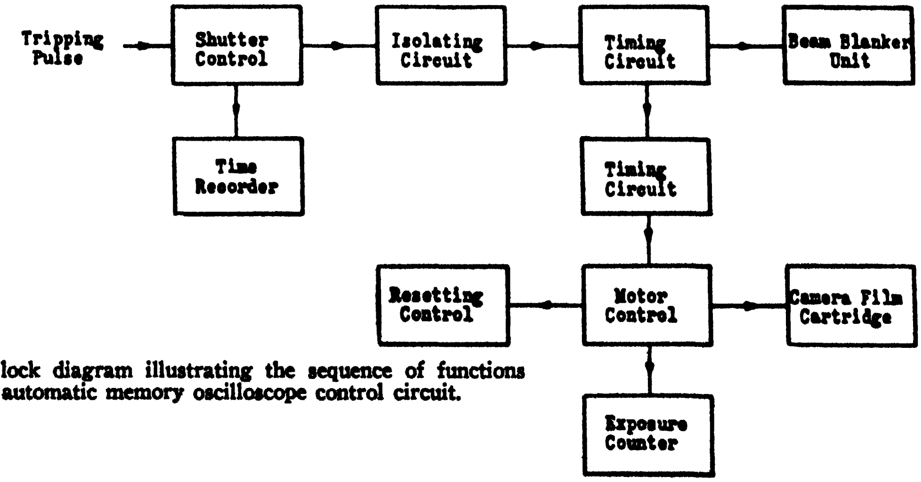


Fig. 19-152.—Block diagram illustrating the sequence of functions performed by the automatic memory oscilloscope control circuit.

Courtesy Electrical Engineering

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[18] Zarem, A. M., "An automatic oscillograph with a memory," *Elec. Eng.*, vol. 65, pp. 150-153, March, 1946.

As the input impedance of the control circuit is quite high, very little current is required for its operation.

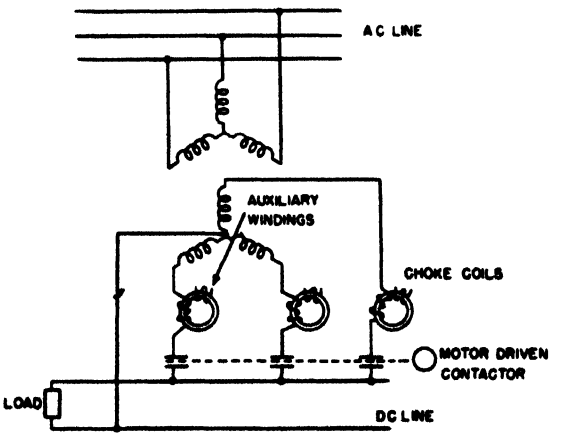
The first operation is a multiple one. The shutter is opened, a clock dial is momentarily illuminated, and an electric pulse is released to an isolating circuit. An adjustable time thereafter, the beam-blanking unit receives its impulse, and a second timing circuit is actuated. The latter operates the motor-control unit, which in turn resets the camera, excites the coil of the step-relay exposure counter, and places the entire control circuit in readiness for the next transient to be recorded. The timing circuits are included in order to allow the adjustment of both memory time and exposure time. The clock mechanism insures that every oscillogram record contains the time at which the initiating transient occurred.

Magnetic Materials Tests[19]

Although magnetic materials possessing substantially rectangular hysteresis loops have been known for some time, only recently has this property been utilized in the design of electric apparatus.

Attempts in the past to construct mechanical rectifiers have been frustrated because of the difficulty encountered in commutation. In order to obtain satisfactory results, the commutation must take place while the current passes through the zero value. In an ordinary a-c wave, the rate of change is the greatest at that point. As a result there is no time for commutation without destructive sparking.

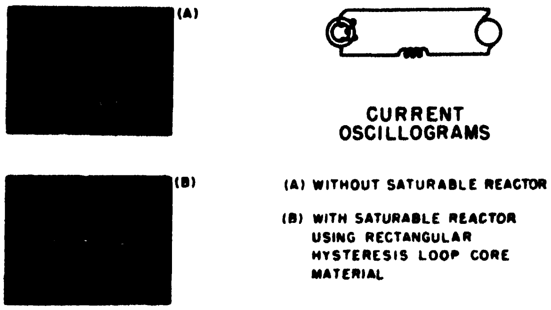
When saturable reactor coils are put in series with the current leads, a step is introduced as the current passes through substantially zero value. This prolongs the time of zero current so that almost sparkless commutation can be obtained. It has been found that far superior results are obtained if these saturable reactor coils have rectangular hysteresis characteristics.



Courtesy Electrical Engineering

Fig. 19-153.—Basic circuit for 3-phase contact rectifier.

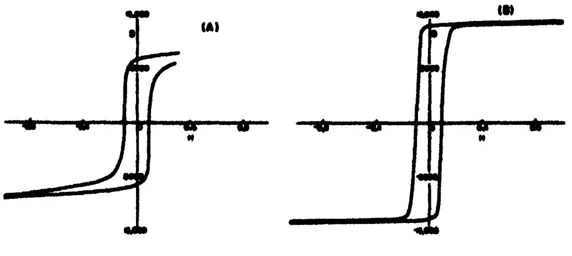
A basic circuit for a 3-phase contact rectifier is shown in Fig. 19-153. The saturable reactor coils are shown together with the windings used to aid in the development of sparkless commutation. Fig. 19-154 shows the load current (A) when the saturable reactors are not in the circuit and (B) when they are included as in Fig. 19-153. In the latter part, the step in the current wave is shown.



Courtesy Electrical Engineering

Fig. 19-154.—Load currents (A) without and (B) with saturable reactors in circuit.

Fig. 19-155 shows the hysteresis loops for two cores of the same material. Core (A) has been heat-treated by standard methods and core (B) is specially processed to give a rectangular hysteresis loop. The commutation takes place at the steep portion of the hysteresis loop. When the width of the loop is small, the current during commutation will remain small but not zero.



Courtesy Electrical Engineering

Fig. 19-155.—Hysteresis loops for two cores of 50 per cent nickel-50 per cent iron alloys having (A) standard annealing, (B) special annealing after drastic reduction.

In order to reduce the current step to zero, an a-c biasing current is supplied through the auxiliary windings. Also, it is desirable that the vertical portions of the hysteresis loops be as steep as possible; and that they approach the saturation value. This extends the commutation interval and minimizes the amount of core material for a given rectifier. Materials which have a rectangular hysteresis loop as shown in Fig. 19-155B fulfill these requirements.

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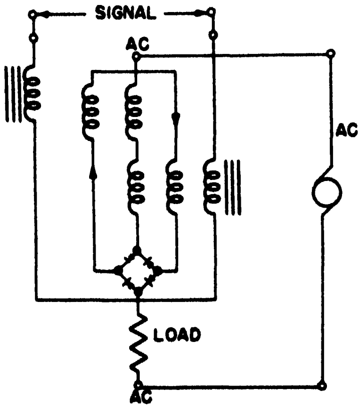
[19] Elmen, G. W., Gaugher, E. A., "Special magnetic alloys and applications," *Elec. Eng.*, vol. 67, pp. 843-845, September, 1948.

Magnetic Amplifiers

Development work at the Naval Ordnance Laboratory has shown that certain advantages in the construction of magnetic amplifiers are obtained by the use of magnetic materials with rectangular hysteresis loops.

One type of magnetic amplifier circuit employing regeneration is shown in Fig. 19-156. Saturable transformers *T1* and *T2* are designed to operate at the knee of the magnetization curve. Because the regeneration is accomplished by the flow of the rectified load current through a feedback winding, a d-c bias is required to minimize the load current at zero signal. When a signal is applied, each transformer is driven into saturation. This increases the magnetizing and load currents.

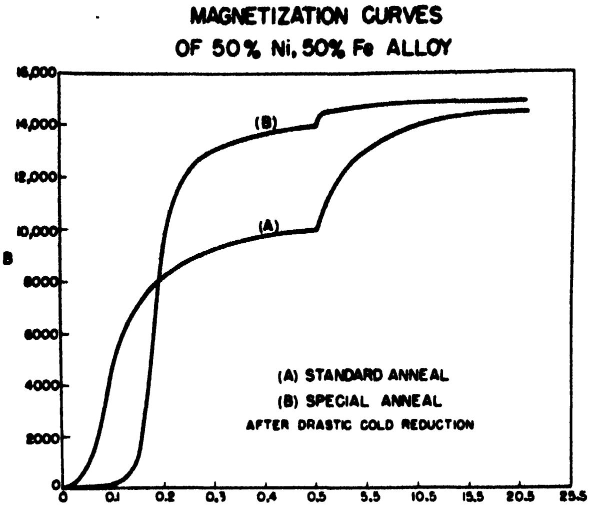


Courtesy Electrical Engineering

Fig. 19-156.—Magnetic amplifier circuit employing regeneration.

From power considerations, it is desirable to operate at high flux density. The higher the density at which the knee occurs, the larger will be the output power capacity of a coil with a core of a given size. Also, in order to increase the amplification, the magnetizing force at the knee should be as small as possible. In a material having a rectangular hysteresis loop, the knee occurs very close to saturation value.

Because of that, the form factor of the load current more nearly approaches that of a sine wave. Fig. 19-157 shows the magnetization curve of two nickel-iron alloy cores of substantially the same composition.



Courtesy Electrical Engineering

Fig. 19-157.—Magnetization curves of two nickel-iron alloys of substantially the same composition: (A) using standard core material; (B) using rectangular hysteresis loop core material.

Core (B) has been given a special treatment to produce a rectangular hysteresis loop while core (A) has been given the standard heat treatment. Fig. 19-158 shows the current waveform for each of these cores, showing improved sinusoidal characteristics of the specially treated material. An amplifier constructed with identical cores of these materials delivered over twice as much power when core (B) was substituted for core (A).

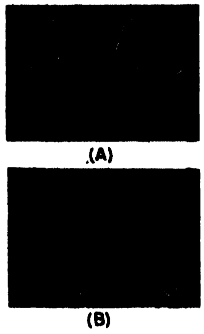


Fig. 19-158.—Current waveforms for the cores of Fig. 19-188, showing improved sinusoidal characteristics of the specially treated material.

Courtesy Electrical Engineering

Microwave Molecular Spectroscopy[20]

It has been known for more than 30 years that very-high-frequency electromagnetic waves would be absorbed by any gas whose molecules possess electric dipoles. The frequencies at which absorption takes place correspond to the rotational frequencies of the molecule. Using war-developed techniques, it is now possible to detect these absorption lines and to measure their frequencies to one part in a million.

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[20] Coles, Donald K., "Microwave molecular spectroscopy," *Elec. Eng.*, vol. 67, pp. 735-741, August, 1948; the following material is based on this article.

The first sharp-line absorption spectrum was obtained by B. Bleaney and R. P. Penrose at Oxford University, England. These investigators introduced ammonia into a resonant cavity, which effectively is a short length of waveguide in which the radiation is reflected back and forth hundreds of times. The effect of the ammonia gas was to damp out the waves and thus to reduce the energy density in the cavity. The change in energy density could be measured by means of a crystal detector.

Bleaney and Penrose found that at pressures below one millimeter of mercury the broad absorption band of ammonia consisted of several series of very sharp lines. This was in general agreement with the prediction of Dennison and others at the University of Michigan. Bleaney and Penrose announced these results early in 1946, and at that time similar results already were being obtained by W. E. Good at the Westinghouse Laboratories and C. H. Townes at the Bell Telephone Laboratories, New York, N. Y.

The American investigators were using a frequency sweep method which presented a small part of the absorption spectrum on the screen of a cathode-ray oscilloscope at any one instant of time (Fig. 19-159). One form of the apparatus used by Good is shown in Fig. 19-160. The microwaves traversed two meters of ammonia gas and then were detected by means of a silicon crystal. The received signal was amplified and applied to the vertical plates of an oscilloscope. The sawtooth voltage which served for the horizontal deflection of the oscilloscope was also used to vary the frequency of the oscillator, so that the horizontal axis of the oscilloscope became a frequency axis. The absorption wavemeter, indicated in Fig. 19-160, is a variable cavity loosely coupled to the waveguide. This type of apparatus has sufficed to detect and study the rotational absorption lines of perhaps a dozen simple molecules in the wavelength region between 5 and 18 millimeters.

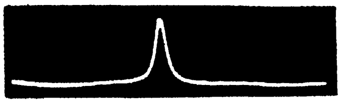
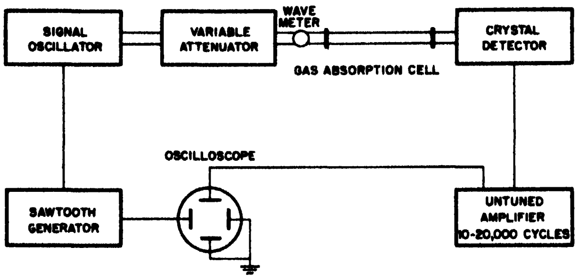


Fig. 19-159.—Presentation of a small part of the absorption spectrum on the cathode-ray-tube screen.

Courtesy Electrical Engineering

A recording would be made at reduced full-wave level and at full-wave level. Usually, these neutral current oscillograms will consist of many frequencies and a mismatch is indicative of a change in the transformer's impedance, such as when a failure occurs.



Courtesy Electrical Engineering

Fig. 19-160.—Block diagram of apparatus used in microwave spectroscopy.

Small changes in impedance result in larger changes in the neutral current oscillograms than occur in the voltage oscillograms taken at the surged terminal. Whether the current neutral method of measurement will detect all failures in a winding obviously depends considerably upon the characteristics of the transformer.

Turn-to-turn failures involving no more than one-tenth of one per cent of the winding have been detected.

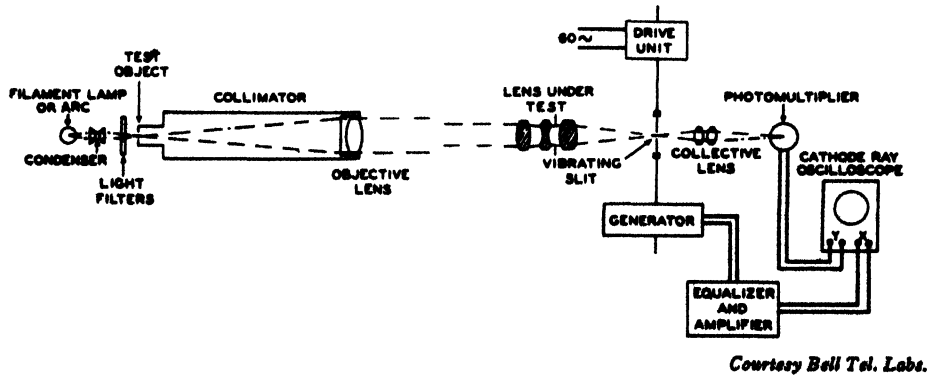


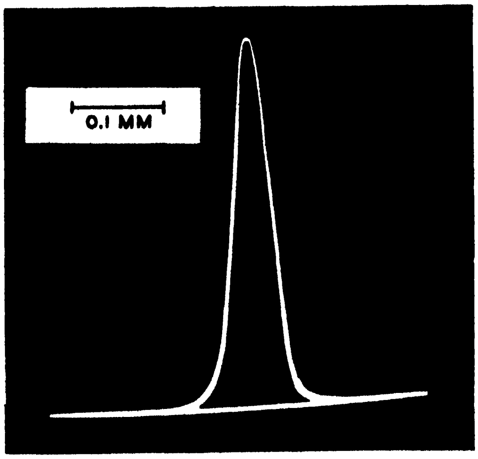
Fig. 19-161. — Basic test setup for lens testing.

Courtesy Bell Tel. Labs.

Lens Testing[21]

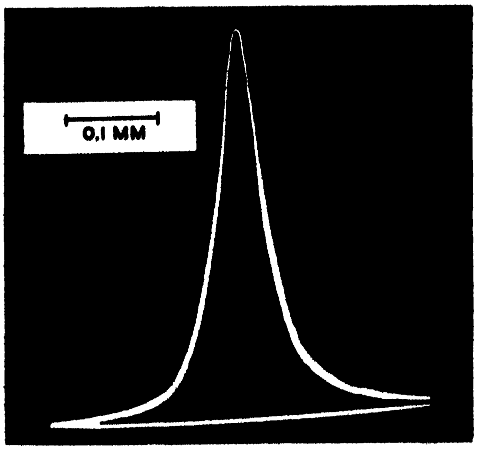
The basic test setup is shown in Fig. 19-161 The optical principles involved are discussed in the original article. Electrically, the hookup is simple and self-evident from the diagram.

The distribution of energy in an image formed by a lens in yellow lights is shown in Fig. 19-162. The distribution of energy in an image formed by a lens in blue light is shown in Fig. 19-163, while Fig. 19-164 shows the light modulation in the image of a grating formed by a lens on axis. Light modulation in the image of a grating formed at 10° of axis is shown in Fig. 19-165. Fig. 19-166 illustrates diffraction spectra present when a coarse grating is placed over the lens under test.



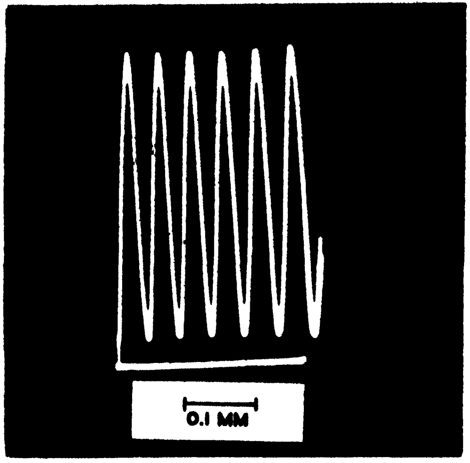
Courtesy Bell Tel. Labs.

Fig. 19-162.—Distribution of energy in an image formed by a lens in yellow light.



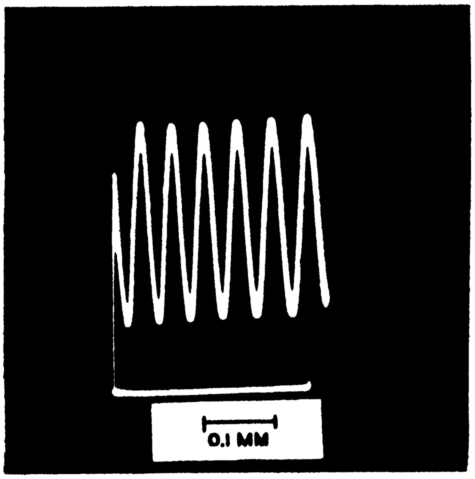
Courtesy Bell System Labs.

Fig. 19-163.—Distribution of energy in an image formed by a lens in blue light.



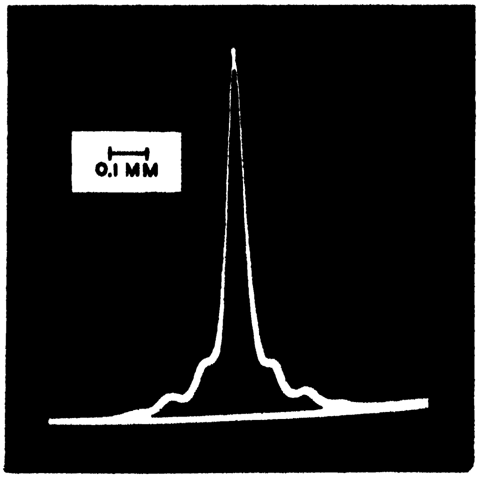
Courtesy Bell System Labs.

Fig. 19-164.—Light modulation in the image of a grating formed by a lens on axis.



Courtesy Bell System Labs.

Fig. 19-165.—Light modulation in the image of a grating formed at 10° of axis is shown.



Courtesy Bell System Labs.

Fig. 19-166.—Diffraction spectra present when a coarse grating is placed over the lens under test.

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[21] Herriot, W., "A photoelectric lens bench," *Jour. Opt. Soc. Amer.*, vol. 37, pp. 472-474, June, 1947.

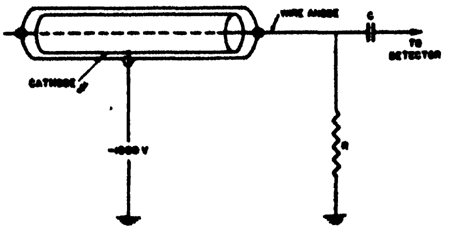
Geiger Counter Tests[22]

Since the first demonstration of the tube counter by Geiger and Muller in July, 1928, the unusual sensitivity of such counters has found widespread applications in the detection of high-speed particles and energetic photons.

A Geiger counter is a gas-filled diode operated in the region of the unstable corona discharge. There are two types of counters, characterized by their filling gases. One uses simple monatomic or diatomic gases, such as hydrogen, air, the rare gases, or mixtures of these, and is known as the nonself-quenching type. The second category includes mixtures of simple gases and small percentages of "quenching" admixtures, which are usually polyatomic organic molecules. In general, the firing characteristics of both types of fillings are very much alike, but the subsequent stages of the discharge and the deionization process are distinctly different.

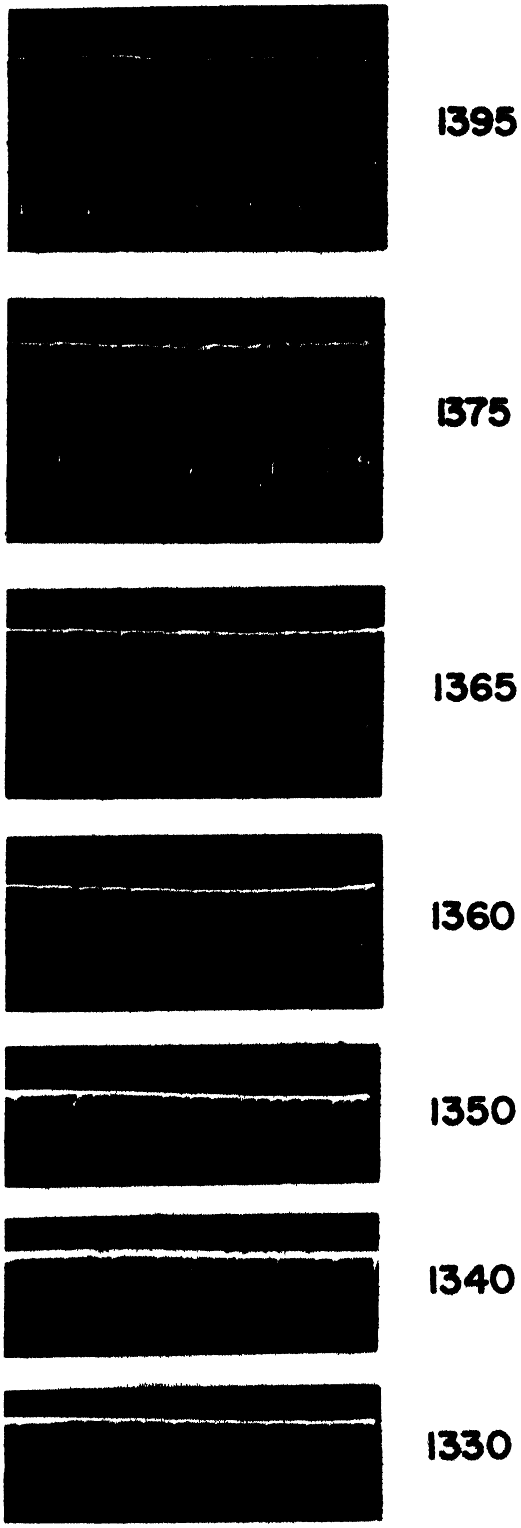
The electrode system of a Geiger counter usually consists of a fine wire and coaxial cylinder. Most tubes are filled with a rare gas combined with a trace of a polyatomic vapor such as alcohol, ether, amyl acetate, and many others. At low voltages, the tube behaves as an ionization chamber with an internal amplification factor of unity. A relatively small potential difference prevents recombination losses, and is sufficient to draw a saturation current from the tube, supplied entirely by the primary ionization. Raising the voltage brings on gas multiplication by impact ionization of the gas molecules.

The multiplication factor increases with increase in voltage and the current delivered by the tube is proportional to the primary ionization up to multiplications of 10^5 or 10^6. Throughout this range, the discharges are single avalanches, each avalanche originating from a primary ion pair and localized within a fraction of a millimeter along the length of the wire.



Courtesy Rev. Sci. Instr.

Fig. 19-167.—Fundamental Geiger counter circuit.



Courtesy Rev. Sci. Instr.

Fig. 19-168.—The transition to Geiger counting is ordinarily very sharply defined. This figure shows the rapid transition from incomplete growth of the discharge characterized by nonuniform pulse heights, to the threshold where each discharge has spread throughout the tube.

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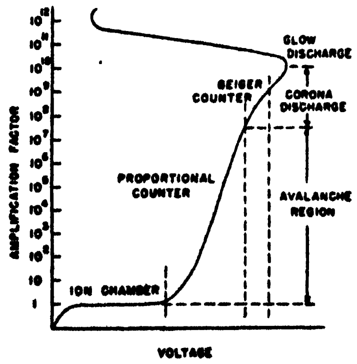
[22] Friedman, Dr., Herbert, "Geiger counter tubes," *Proc. I.R.E.*, vol. 37, p. 791, July, 1949.

At still higher voltages, every avalanche breeds new avalanches, spreading the discharge along the full length of the tube, through the medium of the very-short-wavelength ultraviolet rays generated in each avalanche. The discharge continues to burn until a critical space-charge density of positive ions is reached. The amplification factor then becomes independent of the amount of primary ionization, and all discharge pulses attain equal amplitudes. This condition characterizes the operation of the Geiger counter.

The Geiger-counting threshold is usually determined experimentally by observing, with an oscilloscope, the lowest voltage at which all pulses become equal in size. The fundamental circuit is shown in Fig. 19-167.

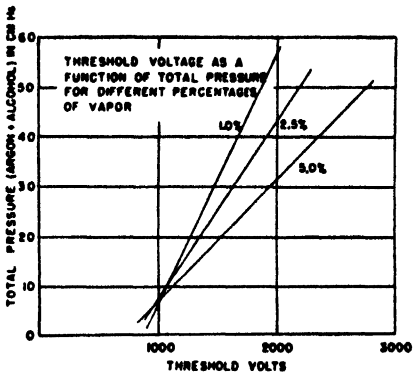
As threshold is approached, statistical fluctuations in the breeding of new avalanches from preceding avalanches may interrupt the chain before the discharge has filled the entire length of the tube. The transition to Geiger counting is ordinarily very sharply defined, as is illustrated in Fig. 19-168, which shows the rapid transition from incomplete growth of the discharge characterized by nonuniform pulse heights, to the threshold where each discharge has spread throughout the tube.

The number of discharges is directly related to the number of primary particles striking the tube, and does not depend appreciably on the applied potential over a range of a few hundred volts known as the "plateau." At higher voltages, the condition of a self-sustained corona is reached and the discharge maintains itself until the potential is removed. Sufficiently high potentials bring on the transition to a glow discharge in which the current rises very rapidly and the voltage across the electrodes falls to a low stable value. The complete voltage characteristic of the cylindrical ionization tube is illustrated in Fig. 19-169. Fig. 19-170 illustrates the effect of argon pressure, percentage of alcohol admixture, and the size of the electrodes on the threshold voltage.



Courtesy Proc. I.R.E.

Fig. 19-169.—Complete voltage characteristic of the cylindrical ionization tube.



Courtesy Proc. I.R.E.

Fig. 19-170.—Effect of argon pressure, percentage of alcohol admixture, and the size of the electrodes on the threshold voltage.

Number Displays[23]

The number is displayed across on the screen of a cathode-ray tube, the characters being directly written by movement of the spot. The effect of a steady display is created by a high-speed repetitive traverse over the outlines required, a black-out being effected between the characters and on the fly-back.

The spot control is achieved by a combination of alternating figure-producing potentials and steady spacing of layout potentials. The former are provided by a common *figure generator* which uses a continuously rotating multiunit variable capacitor to modulate a 500-kc supply, generated by a built-in oscillator. The outputs are rectified to give the *X* and *Y* component potentials of the integers at an audio frequency.

A layout selector connects the tube in sequence to the required figure potentials and adds the steady potentials to locate each in its appropriate position in the frame.

Basic Principles

The technique is based on the cathode-ray tube, the display being directly written by movement of the spot and not by variation in brilliance as in television and other reproduction methods.

Thus to reproduce the figure "2," the spot is caused to traverse the outline of this figure many times a second. Persistence of vision causes the result to appear as a solid unbroken figure. The potentials required to produce this traverse are of an alternating nature. The individual figures are built up into a complete display or frame by superimposing steady potentials, which have the effect of shifting the center of each figure to any required point of the frame. The figures are written one at a time, each with its own combination of alternating and location potentials. The complete trace is repeated several times a second and again persistence of vision creates the appearance of a steady display. This effect is greatly assisted by the afterglow nature of the tube screen.

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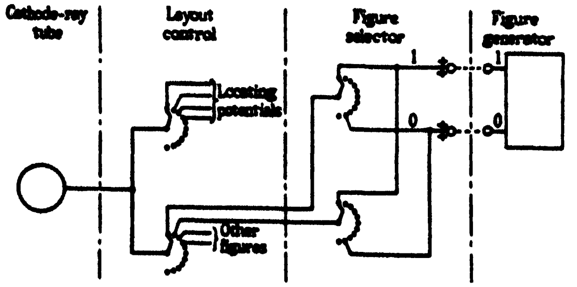
[23] Baker, G. T., "Number display using the cathode-ray tube," *Jour. I.E.E.*, vol. 94, p. 421, November, 1947.

Elements of System

The system has four main components, shown in outline in Fig. 19-171. These are: (a) The cathode-ray tube. This can be any of the standard units depending only on the size and type of the display. (b) The layout control. The face of the tube can be considered as a frame containing a number of divisions in each of which one integer can be displayed; the layout control locates each division. (c) The figure selector. Each division of the frame has an associated figure selector which determines which integer will appear in that division. (d) The figure generator. This provides the alternating potentials for producing traces corresponding to the integers, 1 to 0. One such unit will supply a large number of tubes and thus the system is particularly suited to provide a multiplicity of adjacent but otherwise independent displays.

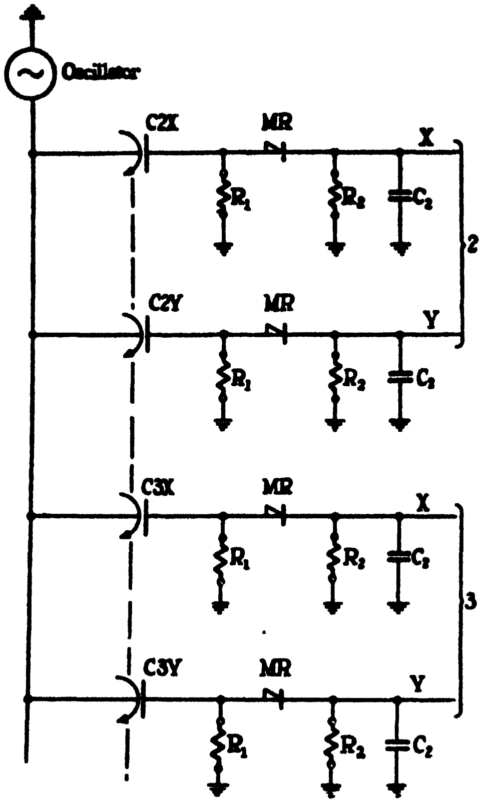


Courtesy Jour. IEE (London)

Fig. 19-171.—Elements of system for number displays.

Figure Generator

The writing potentials for delineating the individual figures are produced by the equipment shown in Fig. 19-172. The oscillator functions at about 500 kc and feeds the common rotor of the multiunit variable capacitor shown as *C2X*, *C2Y*, *C3X*, *C3Y*, etc.



Courtesy Jour. IEE (London)

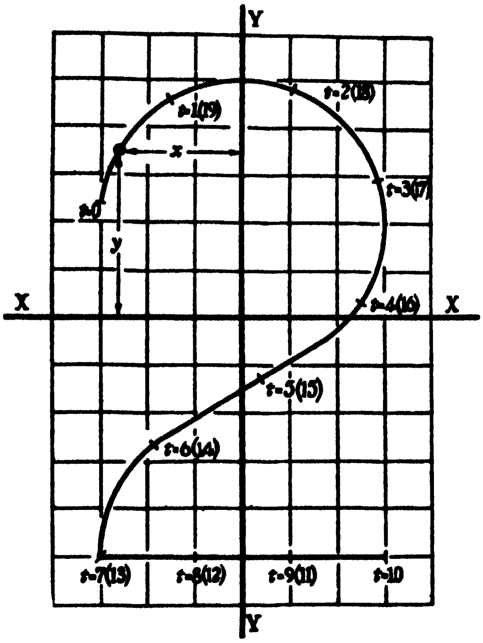
Fig. 19-172.—Writing potentials for delineating the individual figures are produced by this equipment.

The rotor is driven directly from the shaft of a small high-speed motor. The stationary plate of each capacitor unit is connected through a load resistor *R1* to ground, and the varying high-frequency potential developed across this resistance is rectified by the dry-plate rectifier *MR*. The smoothing unit *C2* is introduced to prevent the high-frequency fuzzing the output, which is normally situated in the lower a-f band.

Each figure, with the exception of "1," requires two control potentials, and there are thus nineteen plates and output terminals. It will be seen that the output potenial depends on the instantaneous capacitance of the rotating vanes, and if these vanes are correctly shaped, almost any desired output can be achieved.

Capacitor Vanes Design

Consider the reproduction of an outline, for example, the "2" shown in Fig. 19-173. Imaginary axes can be drawn through the resting position of the spot, one parallel to the deefiction produced by the *X* plates, the other parallel to the *Y*-plate deflection. The deflections are substantially proportional to the applied potentials, so that the distances from these axes, i.e., the



Courtesy Jour. IEE (London)

Fig. 19-173.—Reproduction of an outline.

X and Y co-ordinates, represent the required deflection potentials. The outline shown is, therefore, a relationship between X and Y of the form

$$X = f(Y). \tag{1}$$

This curve is traced out by the movement of a spot, and the relationship is more accurately expressed in terms of a parameter t

$$x = f,(t) \tag{2}$$

$$y = f2(t) \tag{3}$$

where C_x, C_y are the capacitances of the X and Y capacitors respectively.

The traverse is repeated many times a second, hence these represent periodic functions. The load resistance ($R1$, Fig. 19-172) is kept relatively small and the output potentials are roughly proportional to the instantaneous capacitances of the variable capacitors.

$$C_x = f1(t) \tag{4}$$

$$C_y = f2(t) \tag{5}$$

where C_x, C_y are the capacitances of the X and Y capacitors respectively.

The outline should have an even illumination and the spot must move at a constant velocity along the trace. The curve is, therefore, divided into a number of equal divisions, each of which represents a time unit. In practice it has been found more convenient to divide each figure into its salient features and to make each feature an integral number of time units while preserving the over-all equality as near as possible. This makes it easier to reproduce those features which are repeated in two or more figures. Thus, in "2" the baseline is allowed three, and the remainder of the curve seven, out of the ten divisions into which the trace is divided.

The rotor has a constant angular velocity and the time divisions also represent angular positions of the shaft. Equations (4) and (5) then become

$$C_x = f1(\theta) \tag{6}$$

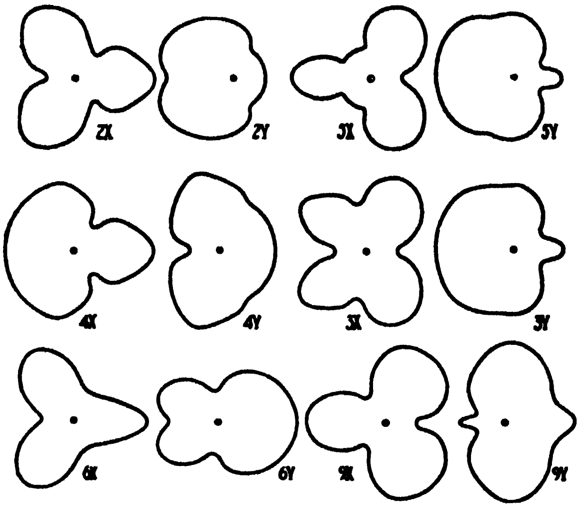
$$C_y = f2(\theta) \tag{7}$$

giving the capacitances required from the X and Y capacitors for any position of the shaft. These two polar curves can be obtained directly from Fig. 19-173 and, if the fixed blades were of negligible width, they would represent the desired rotor shape.

In practice, it is necessary to make the fixed plates as wide as possible and an 18° segment has been found convenient. Before equations (6) and (7) can represent overlapping area of capacitor vane, two corrections are necessary. First, the radial scale must be adjusted so that the divisions represent equal increments of area on the segment. To achieve this, the radial units are made proportional to the square of the distance from the origin. Second, the perimeter does not cut the fixed blades in an arc of a circle, but, in general, includes varying radii. The actual capacitor area is represented by an integral taken between two radii, 18° apart. Fortunately, the sine functions are straight lines respectively, and, therefore, this effect does not entail a great deal of modification, except discontinuities. One or more of these occur in most figures and the reversals at either end of the traverse provide further examples.

A theoretically perfect solution cannot be obtained, but the shapes of the figures are not critical and approximations can be obtained which give a divergence from the ideal almost unnoticeable on the trace. It has been found that once the salient features of the various figure shapes have been studied, a great deal of latitude exists in the plate shape, while still preserving a reasonably symmetrical figure; this latitude can be used to make the plates easier to cut.

Draw the required outline on a Cartesian frame as shown in Fig. 19-173. Divide the trace into a convenient number of equal divisions, and make a polar plot on an equiarea scale, with the semicircle divided up into as many segments as there are trace divisions. The remainder of the circle is used for the reverse traverse. Correct for discontinuities by approximate means and generally smooth the final result to give a simple rounded figure. Typical outlines achieved by the above methods are shown in Fig. 19-174.



Courtesy Jour. IEE (London)

Fig. 19-174.—Typical outlines achieved by methods described in text.

A further correction can be introduced to allow for the fact that the value of the load resistance is not negligible, since electrical efficiency requires this to be fairly high in comparison with the capacitor impedance. The effect, however, is to give a slight flattening at the upper end of the figure, and it is fairly simple to allow for this when drawing the original outline. The required outlines were drawn several times full size and reproduced as a smooth groove on a phenol-fibre board. The latter was then used as a master on a pantograph engraving machine to inscribe the profile at the correct size upon a steel sheet. This was cut out and case-hardened, and it can be used as a template to produce any number of the required vanes which are formed in aluminum on a similar metal or alloy.

Mechanical Details

The principal dimensions were obtained by the following process. A value of 20,000 ohms is used for the load resistance (*R1*, Fig. 19-172) and was suggested by consideration of the output impedance when supplying a number of tubes; too high a value, for *R*, leads to figure distortion. An examination of the potential-divider circuit formed by the reactive modulating capacitors and the load resistor suggested that the minimum impedance of the former must be at least 1.5 times the value of *R*, to avoid undue flattening at the top of the figures.

A frequency of about 500 kc was found most suitable for the oscillator, this value being high enough to keep the capacitors small, without involving the special precautions entailed by the higher radio frequencies. The maximum capacitance of the modulating capacitor was thus fixed at about 10 μf. A plate separation of at least 15-20 mils is essential for ease of construction and maintenance; thus the capacitor requires some 0.75 inch of plate area to give the required capacitance. As previously described, the maximum section that allows a simple layout of vane shape is 18° or 1/20 of a revolution. Internal and external diameters of 1.5 and 7.5 inches, respectively, give a close approximation to the required area.

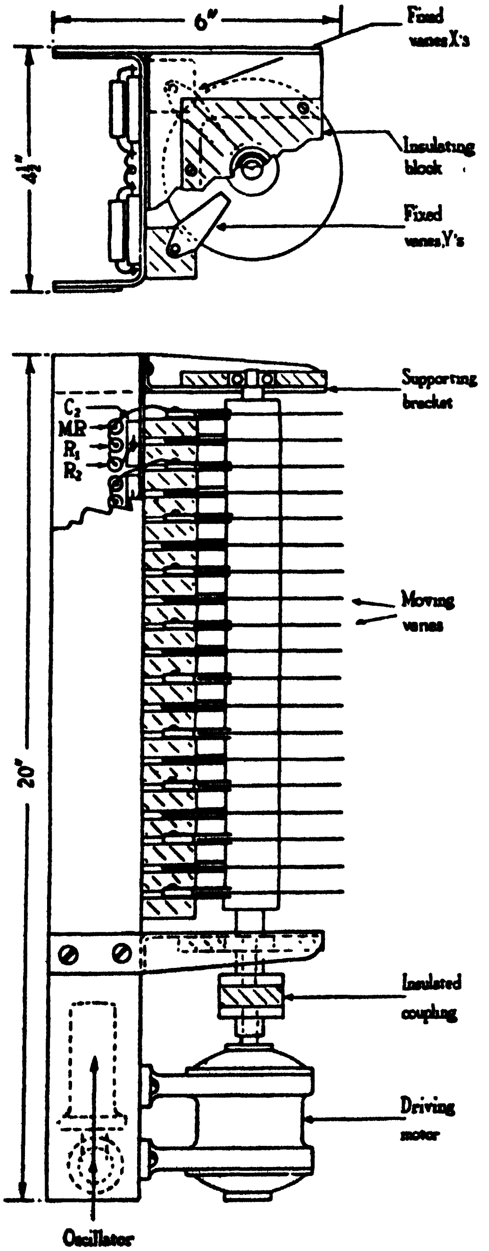
The design, shown in Fig. 19-175, was conditioned by its association with telephone-type equipment, but it is unlikely that another location would demand more than detail modifications. The folded steel chassis is approximately 20 × 4½ inches, the internal recess being just under 2 inches and sufficient to contain all the purely electrical components.

The capacitor shaft is supported by two pressed steel brackets, designed for maximum rigidity. The center face of each bracket is largely cut away, and the space filled by a phenol-fiber square, ¼ inch thick, bolted to the bracket. Roller bearings are simply pressed into the fiber, since this has proved sufficient support for the light stresses involved. The insulating member of the coupling is a disk ½ inch thick, into which studs on the capacitor and motor elements project about 0.1 inch on opposing diameters. By these means the residual capacitance is kept small.

The rotor vanes are assembled on the shaft in pairs, the *X* and *Y* elements of each figure being pinned to the interposed spacer. Thus the components of each digit are maintained in accurate phase relationship, while the unit as a whole can be rotated on the shaft. This makes it possible to adjust the maxima and minima of the various figures so that the load on the oscillator is largely constant.

The fixed blades are mounted on insulating blocks, disposed alternately on either side of the chassis, the fixing holes being somewhat larger than necessary to allow for a certain amount of adjustment before clamping. Further adjustment can be made by slightly bending the fixed blades, this having a marked control over the figure size and shape. The rectifiers (*MR*, Fig. 19-175) and the load resistors (*R1* and *R2*) are situated within the chassis immediately beneath the associated, fixed blades and the wires carrying high frequency are thus extremely short. A cover over the whole unit and smoothing capacitors on the outgoing wires ensure that the high frequency cannot interfere with adjacent apparatus.

A 10-watt motor provides ample driving power, and a synchronous type running at about 3,000 rpm is useful, although a lower speed can be used for some applications. The single-tube oscillator, situated inside the chassis beneath the motor, was designed to work off 100 volts, but very little margin is available at this voltage. Accordingly, in future designs, it is proposed to project both tubes through the chassis in the conventional manner.



Courtesy Jour. IEE (London)

Fig. 19-175.—The mechanical design of number display equipment.

The Control Circuit

A representative circuit for producing a display of numbers is given in Fig. 19-176. The face of the tube can be considered as a frame wherein the integers can be arranged at will by means of the layout control. A convenient method of control uses the telephone-type uniselector, the contacts of which correspond to divisions of the frame.

Each division is located on the screen by a fixed direct potential from one of the uniselector banks, while at the same time, corresponding contacts on other banks connect the appropriate deflection plates to that figure selector which determines the integer to appear in that division. The means used for figure selection depend on the application. For a manually controlled display, the selectors take the form of ten-point hand-set switches. For automatic control, the selection is effected by that part of the controlling apparatus whose condition is to be recorded. The selection in Fig. 19-176 is carried out by means of relay contacts, this being a convenient arrangement, since it corresponds to the circuit conditions encountered in telephony, totalisors, and calculating machinery.

The integers are selected by operating the relays in various combinations, the codes used in Fig. 19-176 being *1-W, 2-X, 3-Y, 4-Z, 5-WX, 6-WY, 7-WZ, 8-XY, 9-YZ, 0-XZ.*

The operation is initiated by connection of the start wire to earth, causing relay *P* to be energized through the magnet interrupter springs and the wipers to move from the home or resting position. In the second position, wiper *1* connects the *X1* plate to one of the *X* terminals of the figure generator, the particular terminal being determined by the combination of relays set up to represent the first digit.

Similarly, wiper *2* connects the *Y1* plate to the corresponding *Y* terminals of the figure generator, and the spot, which is brought up on the tube by a method described later, is driven rapidly around the outline of the required integer. A steady potential is injected from wiper *3* into the *X2* plate such as to locate the integer on the left-hand side of the screen. By using the second *X* plate to control the frame position, the difficulty of combining a steady potential to the tube with a common alternating potential is overcome. Interaction between relay *P* and the uniselector magnet causes the wipers to step to the next position.

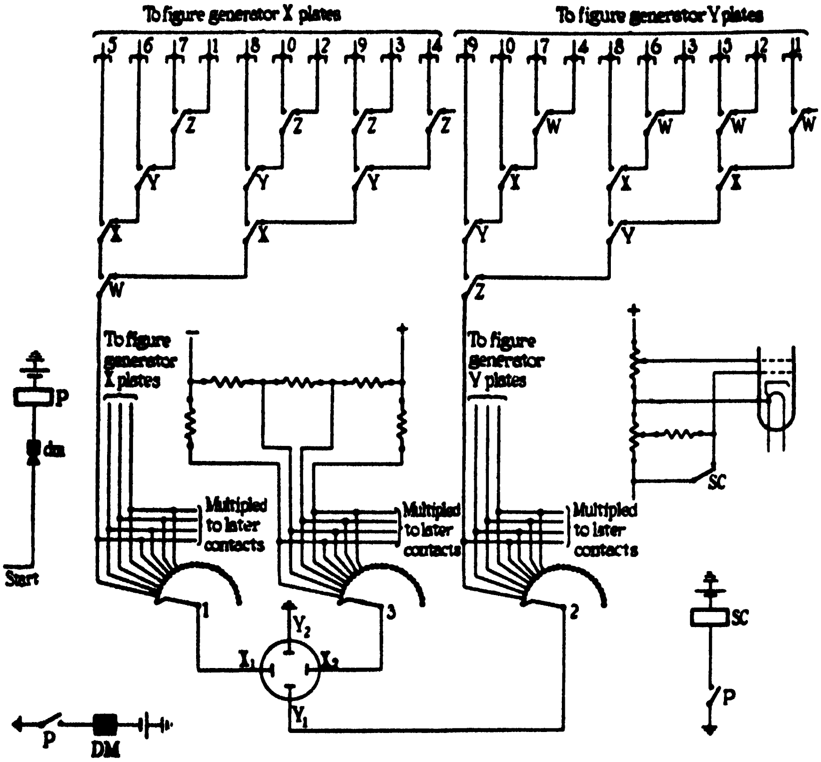

To figure generator X plates
To figure generator Y plates
Z
Y
X
W
To figure generator X plates
To figure generator Y plates
Multipled to later contacts
P
dm
Start
P
DM
SC
1
2
3
Y_2
X_1
X_2
Y_1

Fig. 19-176. — Representative circuit for producing a display of numbers.

Courtesy Jour. IEE (London)

Ultrasonic Tests[24]

The propagation of supersonic waves in bulk rubbers has been studied from 40kc to 10 Mc and from 60° to 60° C.The wave velocity was found to increase with decreasing temperature, leveling off both at high and low temperatures, and was found to increase slightly with frequency.

Peaks in attenuation as a function of either temperature or frequency were observed, the peaks occurring at lower temperatures for lower frequencies. The peaks for butyl, a high loss rubber, are broader and higher than those for GR-S and Hevea, which are lower loss rubbers.

The results are in qualitative agreement with data obtained by strip methods at audio frequencies. However, for bulk waves the real and imaginary parts of two elastic constants, the bulk and shear moduli, determine wave velocity and attenuation; hence, independent measurements of shear wave properties are necessary to evaluate these constants.

At ultrasonic frequencies, measurements of bulk wave transmission can be readily carried out.

Experimental Method

A pulsed compressional wave is generated and received in a liquid by a pair of crystal transducers. By observing the time shift and amplitude change of the received signal when a rubber sample is inserted in the path of the beam, the bulk wave velocity and attenuation can be calculated. The duration of the pulse is such that interference effects do not occur in the first received signal. However, the pulses used are of sufficient length that the group velocity of the pulse may be assumed equal to the phase velocity of continuous waves of the same frequency.

A block diagram of the apparatus is shown in Fig. 19-177. The synchroscope triggers the pulse generator at the same instant that it starts its sweep. A negative pulse from the pulse generator cuts off a tube which normally short circuits the tank of the oscillator. Thus the oscillator is opened for the duration of the pulse. The oscillations are amplified and fed to the transmitting crystal, which generates an interrupted train of waves in the transmitting liquid. The crystals used are of sufficient cross section to insure a well collimated plane wave.

This wave is picked up by the receiving crystal and the resulting signal is amplified and fed into the peak-

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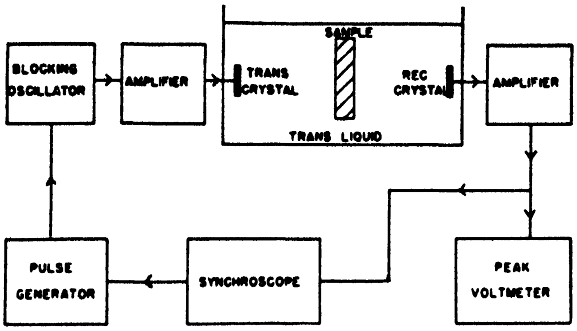
[24] Ivey, Donald G., Mrowca, B. A., and Guth, Eugene, "Propagation of ultrasonic bulk waves in high polymers," *Jour. Appl. Phys.*, vol. 20, p. 486, June, 1949.

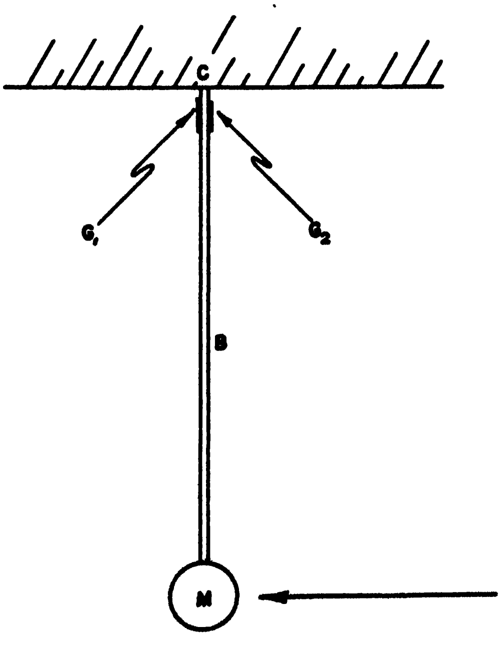


Courtesy Jour. Appl. Phys.

Fig. 19-177.—Block diagram of experimental setup used in ultrasonic propagation tests.

reading voltmeter and the synchroscope. When a rubber sample is inserted in the path of the beam, the wave pattern on the synchroscope screen shifts in position and changes in amplitude. Since a part of the liquid path has been replaced by the sample, the velocity and attenuation in the sample relative to the transmitting liquid can be obtained.

The transmitting liquids used had negligible attenuation compared to that of the rubber, so that the sample attenuation was obtained directly, assuming no reflection at the liquid-sample interfaces. This assumption is valid of liquids of approximately the same acoustic impedance as rubber are used.



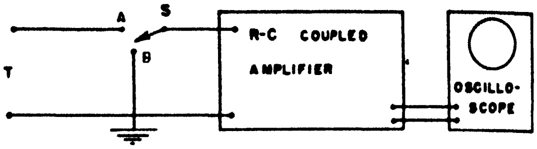
Courtesy Rev. Sci. Instr.

Fig. 19-178.—Simple mechanical vibrating system.

Potential Indicator[25]

The resistance-capacitance coupled amplifier is unsatisfactory for d-c and low a-c amplification. Using a chopper or electronic switch in the input circuit permits satisfactory use of the *R-C* coupled amplifier.

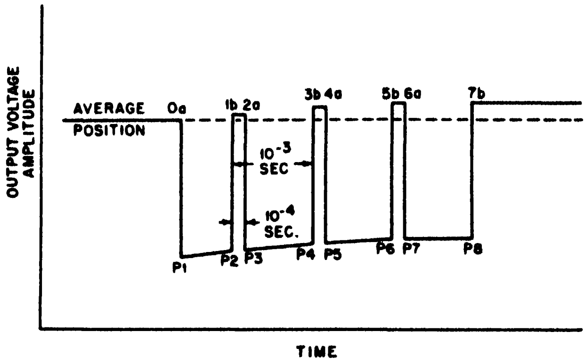
Fig. 19-178 shows a simple mechanical system which, when struck in the direction of the arrow, will contain high-frequency vibrations traveling up and down the band as well as a low-frequency component in the pendulum motion of the mass *M*.



Courtesy Rev. Sci. Instr.

Fig. 19-179.—Basic requirements for the switching process.

Fig. 19-179 is a schematic diagram of input switch mechanism which periodically refers the amplifier input to ground potential. Fig. 19-180 is a hypothetical treatment of step function by input switch mechanism on *R-C* coupled amplifier. Fig. 19-181 is a comparison of hypothetical treatments of complex signal by direct and *R-C* coupled amplifiers. Fig. 19-182 is the schematic circuit of reference potential indicator.



Courtesy Rev. Sci. Instr.

Fig. 19-180.—Step voltage applied across the terminals at *T* in Fig. 19-179.

With reference to Fig. 19-178, the mass *M* is attached to a long, thin bar *B*, which, in turn, is securely clamped at *C*. Strain gages of the fine wire resistance type are cemented at *G1* and *G2*. Suppose now the

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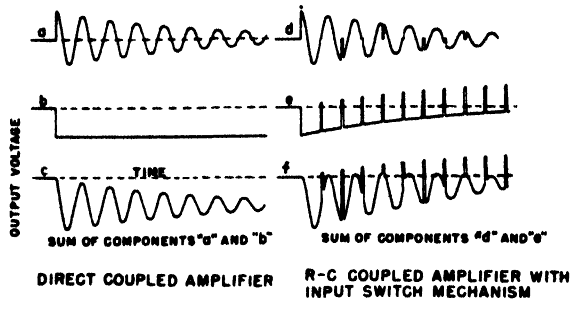
[25] Kammer, E. W., "A constant time interval reference potential indicator for use with *R-C* coupled amplifiers," *Rev. Sci. Instr.*, vol. 17, p. 102, March, 1946.



Courtesy Rev. Sci. Instr.

Fig. 19-181.—Comparison of hypothetical treatments of complex signal by direct and *R-C* coupled amplifiers.

mass *M* receives a sharp blow in the direction indicated by the arrow. Transverse waves composed of many frequencies will travel up the bar, be reflected at *C*, return to *M*, and be reflected again, etc.

The mass *M* will acquire a certain velocity due to the impact and will execute slow, periodic motion similar to a pendulum. The resistance change of the gages *G1* and *G2* will accurately follow the low- as well as the relatively high-frequency changes in strain at the point where the gages are located.

However, an *R-C* coupled amplifier, handling the output of a Wheatstone Bridge in which the gages *G1* and *G2* are connected, may not represent the true magnitude of the dynamic strain by reason of its insensitivity to the low frequency produced by the motion of the pendulum.

The signal is fed through the usual *R-C* coupled amplifier from which the magnitudes of these low-frequency signals can be accurately estimated and at the same time retain all the information contained in the high-frequency components without resorting to direct-coupled amplifiers. This is done by chopping the input signal into regular intervals forming an asymmetric carrier which can be amplified by an *R-C* coupled amplifier. An electronic switch operated by a pulse generator performs the chopping process.

Chopping Sequence

The basic requirements for the switching process can be demonstrated with the aid of Fig. 19-179. A step voltage such as that sketched in Fig. 19-180 is applied across the terminals at *T*. When the switch *S* is in position *A*, the amplifier receives the signal and normal amplification takes place. Let this be the case for one millisecond. Then, the switch is moved to position *B* for 1/10 millisecond, which grounds the amplifier input and no signal is passed.

Referring to Fig. 19-180, at the instant the step voltage is applied to the terminals at *T*, the amplifier output voltage will indicate a level such as point *P1*. However, due to the *R-C* coupling networks in the amplifier, a gradual decay from this value will begin immediately after the point *P1* is reached even though the steady voltage is still applied to the input terminals.

The switch now goes over to position *B*, and, due to a small falling off from the position *P1*, point *1b* will be a little above the average output position. Nevertheless, the total change in output voltage represented by the step *P2* to *1b* is the same as *0a* to *P1*.

Continuing to point *2a* the switch is now returned to position *A*, and the amplifier will repeat the same step it received in position *P1* but will start from the slightly lower zero position at *2a*.

Further decay is experienced from points *P3* to *P4* so that when the switch *S* goes to position *B* again at *3b*, the zero level is still further raised, but the step from *P4* to *3b* is again the same magnitude as *0a* to *P1*, if the d-c input voltage is still the same. If this process continued long enough, say for a second, the output voltage pattern would stabilize about some average position.

When the constant potential is suddenly removed from the input terminals as is indicated at *P8* on Fig. 19-180, the step *P8* to *7a* is the same magnitude as *0a* to *P1*. The amplifier output gradually recovers to the average position indicated by the dotted reference line. Consider next a complex input signal having both high- and low-frequency components present simultaneously. This situation is sketched in Fig. 19-181. Here a step voltage has superposed on it a damped sinusoidal voltage. The switching process would then yield an output voltage as shown by the slotted curve *f* on Fig. 19-181.

Circuit Considerations

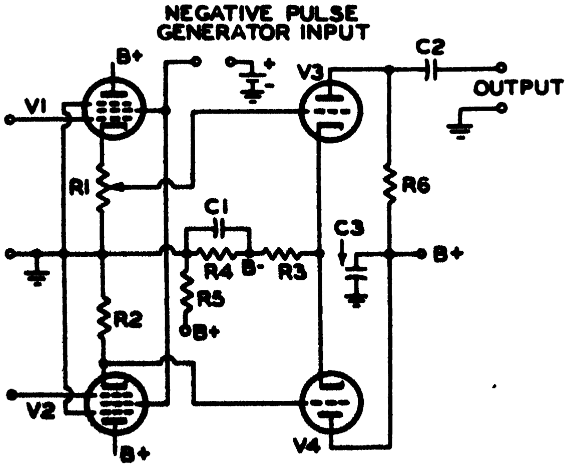
To accomplish the switching required at *S* in Fig. 19-179, an electronic switch was tried, the schematic circuit being shown in Fig. 19-182 Vacuum tubes *V1* and *V2* are operated as cathode followers, while *V3* and *V4* compose a differential amplifier which will accept push-pull signals for amplification but cancels push-push signals.

The differential action of tubes *V3* and *V4* is brought about by a cathode resistor of abnormally high resistance, common to both tubes. Normally, single-ended output is desired and these connections are indicated. On the screen grids of tubes *V1* and *V2* is applied a periodic negative pulse of sufficient amplitude to make these tubes nonconducting. The frequency of these pulses is 1,000 cps, and their duration about 10^{-4} second.

The operation of the circuit is as follows: During the interval between negative pulses on the screen grids, the input signal will appear at the cathodes of *V1* and *V2*. The signal is amplified by *V3* and *V4* and passed on to successive amplifiers and the oscilloscope. When the negative pulse on the screen grids of *V1* and *V2* prevents current from flowing in both tubes, the potential of both cathodes is the same as that of ground, for there is now no voltage drop in resistors *R1* or *R2*. Furthermore, no signal is passed on to tubes *V3* and *V4*.

When tubes *V1* and *V2* are conducting, a bias voltage many times the signal amplitude is present on the cathodes and is reduced to zero during the nonconducting period. This large voltage change would disable the succeeding amplifiers and oscilloscope if permitted to get beyond *V3* and *V4* as a push-push signal and is cancelled.

The circuit as shown can be used with either push-pull or single-ended input. In the latter case, the unused grid is grounded. The pulse generator is synchronized with the output from a 1,000-cps electrically driven tuning fork. In order to keep the corners square on the interrupted signal as viewed on an oscilloscope screen, it is necessary that the amplifiers following the reference potential indicator have adequate high-frequency response. This is not excessively difficult to attain since the amplifiers are required to pass satisfactorily a pulse 10^{-4} second in duration with a repetition rate of 1,000 cps, which is not an unusually severe test.



Courtesy Rev. Sci. Instr.

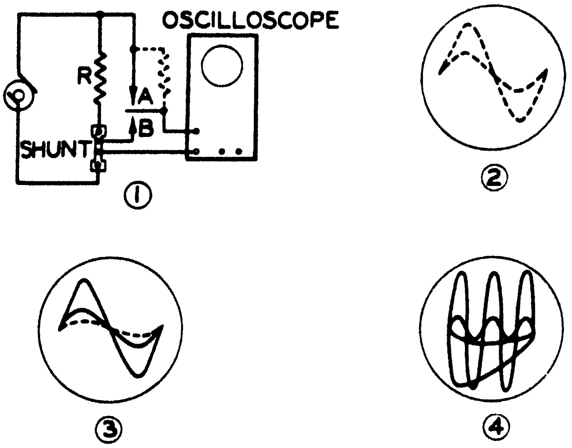
Fig. 19-182.—To accomplish the switching required at *S* in Fig. 19-179, an electronic switch circuit, shown here, was used.

Educational Techniques[26]

The cathode-ray oscilloscope may be used for experimental demonstrations to prove basic radio principles. In a set of experiments devised by RCA for the education of radio servicemen, the oscilloscope has been used successfully to show the effects of tuned circuit adjustments and coupling. This information is necessary to an intelligent understanding of radio servicing. The following series of experiments, eliminating some purely lecture comments, were devised by RCA.

Wave Analysis

In part 1 of Fig. 19-183, an alternator is connected to a resistance load *R*. The voltage across the load is applied to the vertical input circuit of the oscilloscope. A shunt is placed in series with *R* and the signal source to permit current measurements.



Courtesy RCA

Fig. 19-183.—Wave analysis using a vibrator and oscilloscope.

Thus, by means of a single-pole double-throw switch, the oscilloscope can be used as an a-c ammeter or voltmeter. The low side of the oscilloscope is connected to one side of the shunt. With the switch thrown to *B*, the oscilloscope will measure the current flowing through the resistor; at *A*, the voltage across the resistor is measured.

To determine the phase relations between voltage and current, both *E* and *I* waveforms must be received simultaneously. Holding the switch up to *A* to get the voltage wave and then down to *B* to get the current wave, and leaving it in the center as short a time as possible, it would be expected, doing this by hand, slowly, that something like part 2 of Fig. 19-183 would be obtained.

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[26] This material is based on some RCA lecture notes, Series IX, Fall-1935 on the "Cathode ray oscilloscope and its application to the study of tuned and coupled circuits." The material has been edited considerably and revised to bring it up to date so far as possible.

Holding the switch down to *B*, part of the current wave would be obtained. Holding it up to *A*, part of the voltage wave would be obtained. Dotted waves would be expected. Changing the switch arm position and getting these dotted waves, the rate of change must be many times per cycle. Dotted waves would be expected but actually would not be obtained because of the fact that on one cycle there would be a dot, on the second cycle there might be a space, and on the third cycle there might be a dash, etc., so that spaces on one cycle would probably be overlapped by the dots and dashes on the next cycle. It seems, then, that a fill-in would occur to develop a solid waveform line. Assuming a vibrator is used for switching back and forth, at the rate of 78 times per second, using a high-frequency alternator, an overlap will occur to produce a solid line as in part 3 of Fig. 19-183.

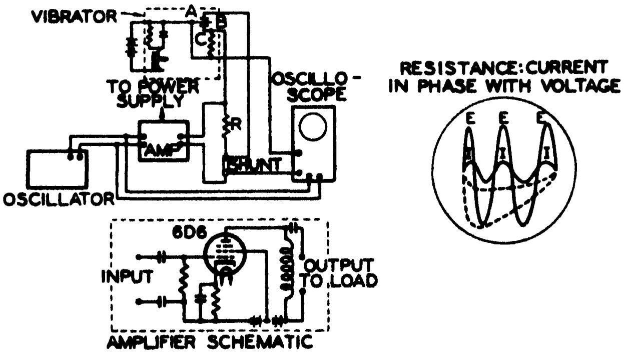
However, another effect must be considered to understand the picture. In part 3 of Fig. 19-183, a third wave is shown dotted, caused by capacitance between contact *A* and the switch arm when the switch arm is in the center position. For high-frequency work, even a small capacitance will serve as a path for current flow. Hence, when the contact of the vibrator is in the center position, it is expected that a solid line will be seen instead of a dotted line, as shown in part 3.

Only the phase relation between voltage and current is of interest, and it would be confusing to have a third wave present on the oscilloscope screen. The vibrator circuit can be modified slightly by adding a high resistance between contact *A* and the movable arm, to short out the capacitance effect and eliminate the third wave at the time when the oscilloscope measures the current wave. When measuring voltage, the resistance is shorted out and does not affect the voltage wave.

Another factor must be considered to interpret the various patterns correctly. A frequency of 45 kc or 45,000 cps is used for the phase-relation tests to show three cycles on the oscilloscope screen. The cathode-ray-tube beam moving horizontally must take one fifteen-thousandth of a second. As the three cycles will be repeated, it is necessary that the beam snap back and retrace the three cycles again. Therefore, the return trace will be seen on the screen. In part 4, the light lines show the snap back of the current and voltage waves.

Demonstration Setup Current and Voltage Phase Relation for Resistance

Fig. 19-184 shows the demonstration setup used to show simultaneously current and voltage waves. In the upper portion of this figure is an auto radio type vibrator. At the left is an r-f oscillator, with a 45-kc output. Therefore, we can show three cycles. A coupling amplifier stage may be used, as shown in the lower part of the drawing, to build up the r-f output to a higher level. The amplifier output is impedance-coupled to the load *R*. The lower lead from the shunt is connected to the low side of the vertical input of the oscilloscope. The high side of the oscilloscope is connected to the center contact of the vibrator. When the vibrator arm *B* moves to *A*, the oscilloscope will show current wave. When the vibrator arm *B* moves to *C*, the oscilloscope will show the voltage wave.

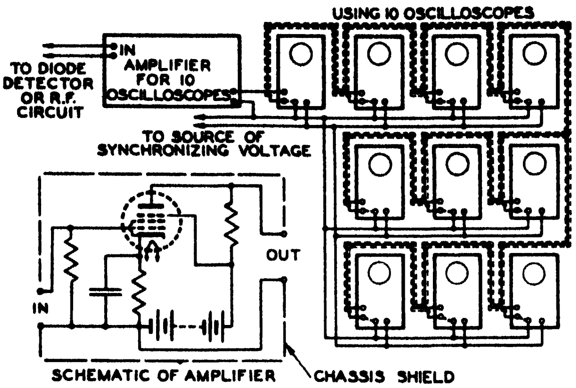


Courtesy RCA

Fig. 19-184.—Demonstration setup used to show, simultaneously, current and voltage waves on the oscilloscope.

Use of Ten Oscilloscopes

Fig. 19-185 shows a hookup of ten oscilloscopes. To get sufficient voltage for operation, a special amplifier is necessary. This amplifier is resistance-coupled, using a pentode power tube. Shielded leads are used to prevent picking up hum. For an r-f demonstration, the grounding of the shielding is removed to prevent the r.f. from disappearing. Hum pickup is then removed by other means such as placing an r-f choke and capacitor between the high side of the line and ground. A special r-f transmission-line system could be developed to serve the same purpose. The source of r-f power has a frequency of 45 kc.


USING 10 OSCILLOSCOPES
TO DIODE DETECTOR OR R.F. CIRCUIT
IN
AMPLIFIER FOR 10 OSCILLOSCOPES
TO SOURCE OF SYNCHRONIZING VOLTAGE
OUT
IN
SCHEMATIC OF AMPLIFIER
CHASSIS SHIELD

Courtesy RCA

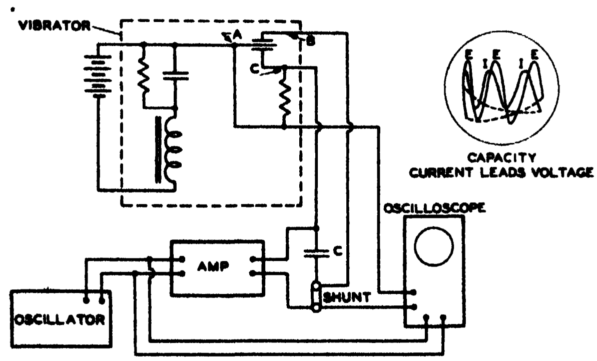
Fig. 19-185.—Hook-up of ten oscilloscopes for mass electronic demonstrations.

Demonstration No. 1

The larger of the two waves is the voltage wave. The smaller wave is the current wave and the retrace lines are caused by the snap back of the "sawtooth" oscillator. The voltage-wave maximum occurs simultaneously with the current-wave maximum. Since the maximum values occur at the same time, the voltage and current is said to be "in phase." In a capacitive circuit, the current leads the voltage by 90°.

Demonstration Setup Current and Voltage Phase Relation for a Capacitance

In this circuit a capacitor *C* replaces *R*, but the remainder of the previous circuit is used. The demonstration circuit is shown in Fig. 19-186.



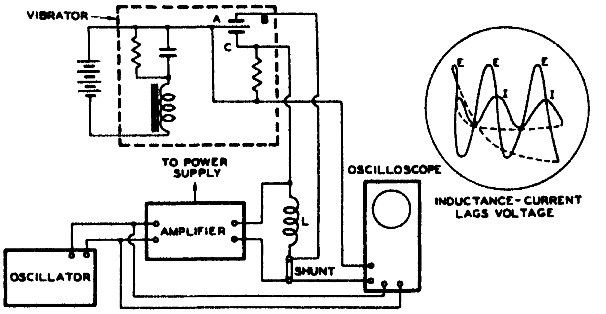
Courtesy RCA

Fig. 19-186.—Demonstration setup for showing current-voltage relations in a capacitive circuit.

Demonstration No. 2

The voltage wave is the larger of the waves. The current wave is the smaller of the two waves. The retrace lines are caused by the snapping back of the "sawtooth" wave used for the timing axis. The current is a positive maximum at this point and the voltage a positive maximum approximately 90° behind the current. The zero line is considered to extend through the center of this figure. It is seen that the current is a maximum when the voltage is zero. Time is in the direction from left to right; it increases as the cathode-ray-tube spot moves left to right. The current leads the voltage by 90°.

The phase relations of current and voltage in a resistance and in a capacitance have been demonstrated. Next, we may show the inductive effect. In the case of an ideal inductance, the current lags the voltage by 90°. In a practical case, the current lags the voltage by something less than 90° due to inherent resistance.



Courtesy RCA

Fig. 19-187.—Demonstration setup for showing current-voltage relations in an inductive circuit.

Demonstration Setup Current and Voltage Phase Relations for an Inductance

The same setup is used as before, except that inductance *L* is substituted for capacitance *C*. See Fig. 19-187. Again two waves are portrayed on the cathode-ray-tube screen — the voltage wave and the current wave. The voltage wave is the larger.

The inductance chosen has very low resistance, so that the current lags the voltage by almost 90°.

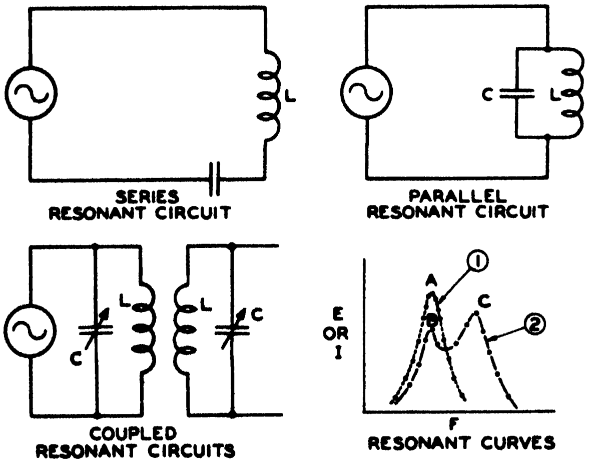
Demonstration No. 3

The voltage is a maximum when the current is zero. It is evident that the current lags the voltage in an inductance by nearly 90°. The resistance of the circuit may now be increased to note the effect on the oscilloscope. A 5,000-ohm resistance is connected in series with the inductive component. The current maximum is reached much sooner than before and the current lags the voltage by approximately 45°.

The phase relations for resistance, capacitance, and inductance have been demonstrated.

Resonant Circuits

In the upper left of Fig. 19-188 is shown an a-c generator connected in series with an inductance *L* and capacitance *C*. At the upper right is an alternator in parallel with an inductance *L* and a capacitance *C*. The circuits are respectively series- and parallel-resonant types. In the series-resonant circuit, maximum current is drawn from the generator at resonance. In the parallel type, minimum current is drawn as resonance. Both types of resonance are used in radio receivers, both in r-f and a-f circuits. The coupled circuits in the lower left of the figure can be considered to be a combination of the two. If we consider the voltage induced by the primary in the secondary to be in series with the secondary inductance, the secondary circuit is considered a series circuit. Since the alternator is connected in parallel with the primary, the primary circuit is a parallel-resonant circuit type.



Courtesy RCA

Fig. 19-188.—Resonant circuit analysis.

If resistance is present in these circuits, they will still be resonant circuits but will act differently than they would in the absence of resistive properties. The behavior of the circuits with reference to frequency may be determined by measurement of the *current* in the series-resonant circuit, the *voltage* across the parallel-resonant circuit, or the *voltage* across the secondary of a coupled circuit, as the frequency of the signal source is varied above and below the resonant frequency.

Parallel Resonance

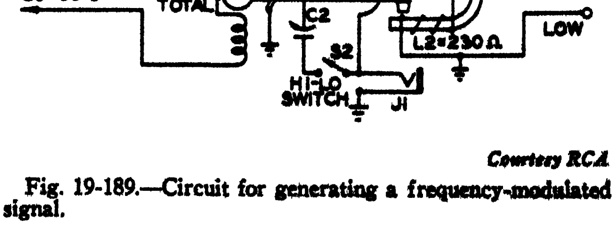
In the lower right of Fig. 19-188 curve *1* is plotted to show the variation in voltage as a function of frequency. Taking a sufficient number of points and connecting them, the curve is seen to increase to a maximum point, such as *A*. The frequency at which maximum voltage is developed, point *A*, is the resonant frequency. Many points are shown on the curve. Taking accurate readings for each one of the frequencies, determining the voltage, and plotting the curves by hand, would be laborious. Therefore, the oscilloscope is more efficient.

Curve *2* represents a resonant curve of a coupled circuit with the secondary tuned to a frequency such as *C*, which is different from the primary resonant frequency *B*. This type of curve may be encountered in alignment of radio receivers. To trace out the response curve automatically, a frequency-modulated generator is necessary.

Description of a Frequency Modulator

In Fig. 19-189 an f-m generator system is shown. To the left we see an a-c motor which is used to drive a shaft. On this shaft are two rotating elements; one is the rotating capacitor used as a sweep capacitor, and the other is an impulse generator. The rotating capacitor is made in two sections so that one section alone can be used, or both sections can be used in parallel. More modern electronic frequency-modulated equipment could be used, but the fundamentals involved here should be understood since old principles and ideas have a way of cropping up again and again in revised form in new electronic equipment. Also, the electromechanical form of frequency modulator is well adapted to low-frequency operation, which was used in these experiments, perhaps more so than the higher operating frequency apparatus in common use today. If the equipment is available, a set of high-frequency experiments could be devised. At this writing, sufficient time is not available to carry forward such work, but it probably will be done in the future.

But, to return to the description of this particular equipment, as the shaft rotates, the capacitance changes from a low value to a high value and from a high value to a low value. If a cable is connected from the jack on



Courtesy RCA

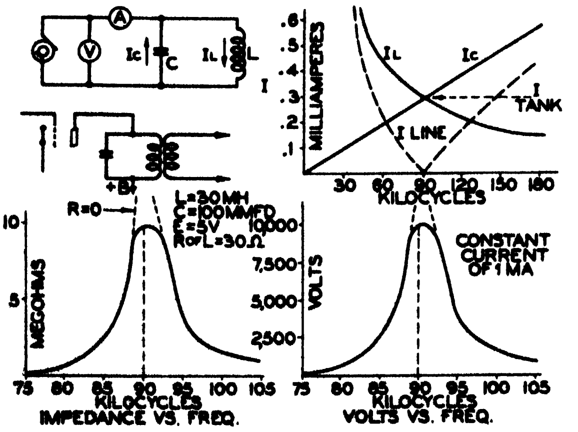
Fig. 19-189.—Circuit for generating a frequency-modulated signal.

the frequency modulator to the jack provided on the signal generator, the oscillator frequency is shifted from a high value to a low value and from a low value to a high value. As the capacitors rotate, the frequency of the oscillator is made to vary, first in one direction and then in the other. This is termed the reverse sweep method.

On the right in Fig. 19-189 is an impulse generator. On the shaft there are two poles, which, when rotating, pass through the magnetic field of the electromagnet. When the poles are across the field, the air gap is small and a large amount of flux is present. When the poles are crosswise, there is little flux. Therefore, as these poles are rotated, the flux will change in amount and, in so doing, will link coils *L1* and *L2,* inducing a voltage in them, of the same frequency as the rotation of the sweep capacitors. This voltage is used to interlock the "sawtooth" oscillator with the sweeping capacitor, so that for each revolution the "sawtooth' oscillator can be kept in synchronism with the frequency modulation.

Parallel Resonance

At the top left in Fig. 19-190 is seen an idealized parallel-resonant circuit. Immediately below it is depicted the practical case of a parallel-resonant circuit driven by a tube which may be considered as a generator of an alternating voltage. In a practical circuit, such as this, it is of interest to know how the value of the voltage delivered by the secondary of the transformer varies with frequency. To do this, we must, in some way, determine the voltage built up across the primary circuit of the transformer. The plate-circuit load for the tube is a parallel-resonant circuit.



Courtesy RCA

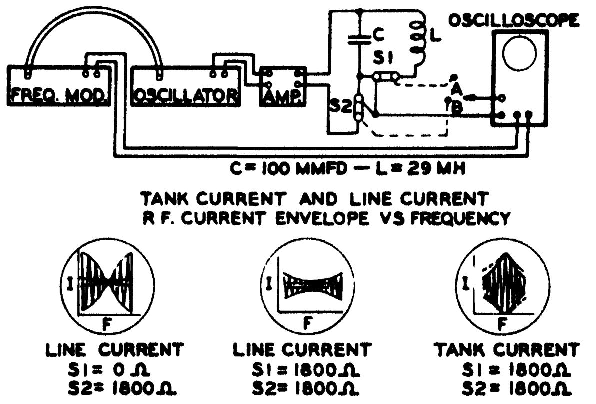
Fig. 19-190.—Analysis of parallel resonant circuits.

We can determine the values for the resonant frequency (90 kc) by first discovering how the currents vary through the capacitive and inductive branches and the value of the total line current at resonance. Calculations may be eliminated by securing a visual representation with the oscilloscope.

Demonstration Setup — Parallel Resonance — Line Current Versus Tank Current

A low frequency is used to avoid oscillating or detuning effects. In Fig. 19-191, the test setup is illustrated. A frequency-modulated generator is required. It is of interest to examine oscillographically the line and tank currents. Shunts are used to permit measurement of the line and tank current. The r-f output wave of the circuit can be checked with the oscilloscope.

First the line current, with no shunt in the tank circuit, may be examined, and then further experiments can be conducted.



Courtesy RCA

Fig. 19-191.—Setup for demonstrating parallel resonance and line current versus tank current.

Demonstration No. 4

At resonance, the line current approaches zero, and at off-resonance the line current is larger. This ties in with the theoretical curve. The line current may be viewed with the shunt in the circuit. The line current has increased. The shunt, a resistor, causes losses which the line must supply. With the resistor in series with the inductance, a slight detuning of the circuit occurs. This effect will be further investigated later on.

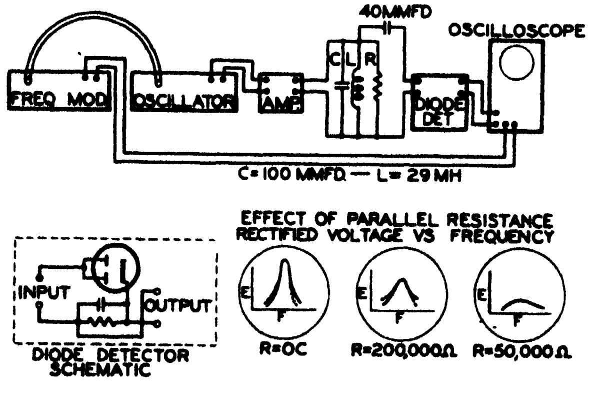
Demonstration No. 5

The tank current is very large at resonance compared to the line current. Theory, therefore, is corroborated without the necessity of a detailed mathematical analysis. It is evident that, in this setup, we may have a signal source supplying a parallel-resonant circuit with a current of only 1 ma, but inside the *L-C* circuit, circulating between inductance and capacitance, the current can reach a much larger value of 20 ma. This is a resonant step-up effect characteristic of an *L-C* circuit of the type shown.

Demonstration Setup — Parallel Resonance — Effect of Parallel Resistance

In a previous figure, the resonant curve of a parallel-resonant circuit was shown. This was a calculated curve and showed the variation of the voltage across the circuit as a function of frequency. Considerable time is necessary to calculate this particular curve. Using the oscilloscope, the curve may be derived easily.

In Fig. 19-192 is a circuit arrangement similar to that used to demonstrate line current and tank current in a parallel-resonant circuit. The output of the frequency-modulated oscillator is fed through the same amplifier that was used before, into a parallel-resonant circuit made up of capacitor *C* and inductance *L*, with resistance *R* in parallel. The voltage across this parallel-resonant circuit is coupled to a diode detector for demodulation of the r-f signal to an audio signal.



Courtesy RCA

Fig. 19-192.—Circuit for demonstration of parallel resistance loading effect.

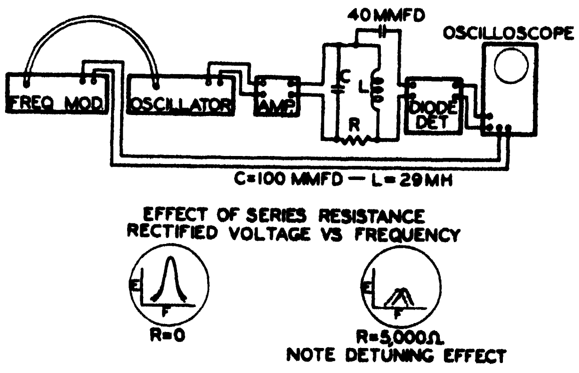
This signal is normally used in alignment of the r-f or i-f circuits of a radio receiver. With no resistance in parallel, the resonant curve is fairly sharp, and a high amplitude is obtained. These curves are taken with the reverse sweep method; that is, the oscilloscope plots one curve with increasing frequency from a low to a high value, and then a second curve from a high value of frequency back down to a low value. At resonance, these two curves are superimposed on each other. The presence of the second curve is indicated by a dotted line on the oscilloscope screen, two lines at each side of the base, as a resonant curve is never quite symmetrical on both sides of resonance.

A 200,000-ohm resistor may be placed in parallel with the resonant circuit to determine the effect of circuit loading. It is seen that the voltage across the circuit is decreased, and while the base of the curve is in the same position as before, the curve is much broader than the previous one. The effect of resistance across a parallel-resonant circuit is to decrease the gain and selectivity.

Placing a lower resistance of 50,000 ohms in parallel with the *L-C* circuit, it is seen that the voltage across the circuit is extremely low and the response curve is also very much broader. In all the various tests with parallel resistance shunts, it is evident that the frequency of resonance has not changed.

Demonstration Setup — Parallel Resonance with Series Resistance

In Fig. 19-193 the setup is similar to that used previously except that *R* is in series with the *L* and *C* components of the parallel-resonant circuit.



Courtesy RCA

Fig. 19-193.—Circuit for demonstrating effect of resistance in series with a tuned circuit.

Demonstration No. 7

In this demonstration, it is noted that there is quite a sharp curve of high value, indicative of good selectivity and good gain. Placing in this circuit a series resistance of 5,000 ohms, the resonant curve value decreases and it is apparent that the reverse sweep curves are not coincident. Some detuning is evident. In the demonstration of line and tank current effects, retuning of the circuit was necessary because of the insertion of series resistance in the parallel *L-C* circuit.

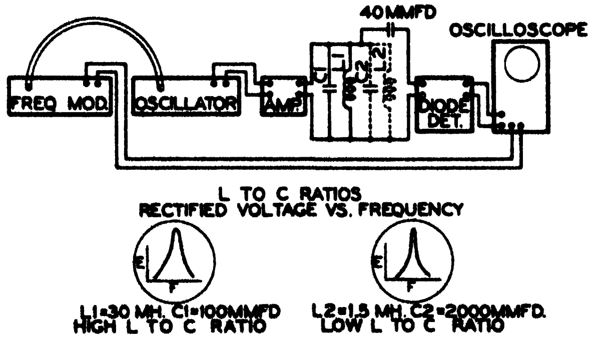
To show the detuning a little better, it is necessary to increase the gain and thus obtain a larger and more easily visible pattern on the cathode-ray-tube screen. We can retune *C* and make the reverse sweep curves coincide. The amount of detuning in terms of frequency can be evaluated by resetting the oscillator and noting the departure from the original frequency value, rather than simply readjusting *C,* which would not of itself indicate the frequency change.

Basically, it is important to recall that when series *R* is placed in a parallel-resonant circuit, the frequency of resonance is affected. The gain is lowered and the selectivity reduced.

Demonstration Setup — Parallel Resonance L-C Ratios

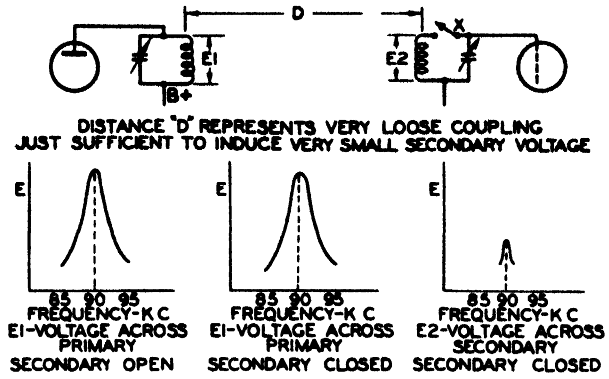
In a parallel-resonant circuit, a change in the *L-C* ratio has a definite effect on the resonant response curve. The test setup is shown in Fig. 19-194. This demonstration setup is similar to the previous one except that a parallel-resonant circuit with one combination of capacitance and inductance, *C1* and *L1,* and another combination using different values, *C2* and *L2,* is used. In the case of the first combination, the inductance is relatively large and the capacitance small. In fact, the inductance is 30 millihenrys and the capacitance is 100 $\mu\mu f$. This provides a high *L-C* ratio. In the second combination, an inductance of 1.5 millihenrys and capacitance of 2,000 $\mu\mu f$ is used to obtain a low *L-C* ratio.

Both of these combinations cause the parallel-resonant circuit to resonate in the vicinity of 90 kc.



Courtesy RCA

Fig. 19-194.—Test setup for demonstrating the effect of the *L-C* ratio on the response of a parallel tuned circuit.


D
X
E1
E2
B+
DISTANCE "D" REPRESENTS VERY LOOSE COUPLING JUST SUFFICIENT TO INDUCE VERY SMALL SECONDARY VOLTAGE
E
85 90 95
FREQUENCY-K C
E1-VOLTAGE ACROSS PRIMARY
SECONDARY OPEN
E
85 90 95
FREQUENCY-K C
E1-VOLTAGE ACROSS PRIMARY
SECONDARY CLOSED
E
85 90 95
FREQUENCY-K C
E2-VOLTAGE ACROSS SECONDARY
SECONDARY CLOSED

Courtesy RCA

Fig. 19-195.—Arrangement for testing coupled circuits.

Demonstration No. 8

The resonant curve is shown here. The broadness of the curve is evident so that we can compare it to the low *L-C* ratio resonant curve.

The resonant curves for the two *L-C* ratios are shown in the lower part of Fig. 19-194. Using the high *L-C* ratio circuit, the selectivity curve is relatively broad, but the voltage at resonance is comparatively high. A narrower curve, more peaked, indicating greater selectivity, is obtained with a higher value of capacitance and lower *L-C* ratio. However, the increased selectivity is accompanied by lower gain and a reduced rise of voltage at resonance.

In the drawing, the reduced gain of the circuit is compensated for by raising the vertical gain of the oscilloscope, so that both curves have approximately the same height; nevertheless, the effect noted above with respect to loss of voltage as the *L-C* ratio is made less should be kept in mind.

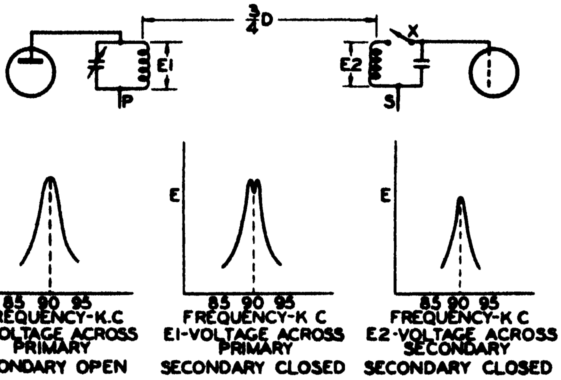
The test setup for coupled circuits is shown in Fig. 19-195 The primary circuit is resonated at 90 kc, and the secondary is also tuned to resonate at 90 kc. In the first experiment, the two coils are separated by a great distance *D.* A switch *X,* in the secondary, is so arranged that the secondary circuit can be opened. When it is opened, the secondary circuit cannot draw energy, a condition equivalent to being withdrawn entirely from the circuit. The primary circuit may then be checked with the secondary circuit either open or closed, as desired, for experimental tests.

With the secondary circuit open, the primary resonant curve is an ordinary one. The primary current causes the induction of a voltage in the secondary. This voltage has the characteristics of the primary voltage but is smaller in magnitude. With the switch in the secondary circuit closed, the voltage developed across the primary remains unchanged because the secondary is remotely situated with respect to the primary. Therefore, only a small amount of energy can be reflected back to the primary.

Viewing the secondary voltage, it is seen that it is much smaller than the primary potential. The sharpness of the response curve is determined by the constants of the secondary circuit. The induced voltage is very small because of the extremely loose coupling between the primary and secondary coils.

Coupled Resonant Circuits — Voltages in Loosely Coupled Circuits

The test setup is shown in Fig. 19-196. If the distance between the coils is decreased, the coupling is increased and a larger voltage is induced in the secondary circuit. A greater amount of energy is transferred from primary to secondary. As the secondary is tuned to 90 kc, maximum transfer occurs at that frequency.

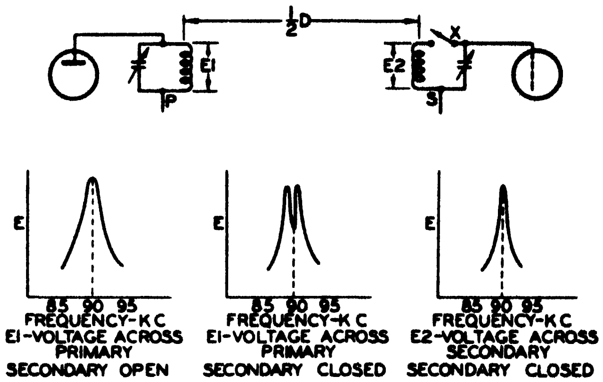


Courtesy RCA

Fig. 19-196.—Arrangement for testing loosely coupled circuits.

If the secondary circuit switch is now opened, and current does not flow in the secondary coil, there is no flux linkage back on the primary, and the primary *L-C* circuit takes on the characteristics of a simple combination of parallel *L* and *C*.

When the switch is closed, the reaction of the secondary on the primary is indicated by a dip in the response curve, and this is shown in Fig. 19-196. The secondary response pattern does not show a dip effect because the *Q* is great enough to compensate for the primary circuit change.

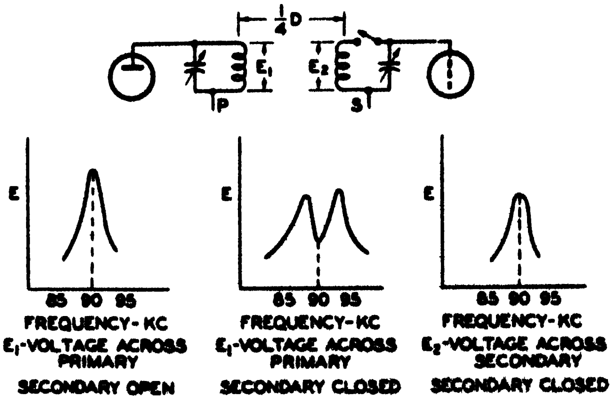


Courtesy RCA

Fig. 19-197.—Arrangement for testing medium-coupled circuits.

Coupled Resonant Circuits — Voltages in Medium-Coupled Circuits

The test circuit is shown in Fig. 19-197. If the primary and secondary coils are brought closer together so that they are only ½ as far apart as they were in the previous example, the amount of energy delivered to the secondary will be greatly increased, and so the effect of the secondary on the primary is very greatly increased. If the secondary switch is opened, the reaction of secondary on primary is virtually nil. If the switch is closed, however, the primary circuit resonant curve is now a letter "M" with a deep valley. The secondary resonant curve is still a smooth resonant curve without the valley in it because of the fact that the sharpness of resonance of the constants of the secondary circuit more than compensates for the difference occurring in the voltage induced in the secondary circuit.

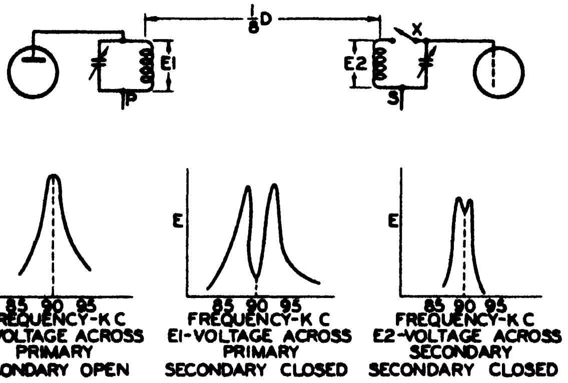


Courtesy RCA

Fig. 19-198.—Arrangement for testing optimum-coupled circuits.

Coupled Resonant Circuits — Voltages in Optimum-Coupled Circuits

As the coils are brought still closer together, to ¼ of the original distance apart, the primary circuit voltage, with the secondary open, is, of course, the same as it was before. See Fig. 19-198. When the switch is closed, the primary resonant curve has a very much deeper gash cut in it than it had previously and it is also broadened. Viewing the secondary curve, it is noted that the effect of the dip of the primary on the secondary is compensated for by the secondary circuit characteristics. However, the compensation is not complete. The secondary curve is beginning to flatten out at the resonant point. When this flattening begins to occur, the coupling at which it occurs is called "optimum coupling."


⅛D
X
E1
E2
P
S
E
85 90 95
FREQUENCY-KC
E1-VOLTAGE ACROSS PRIMARY SECONDARY OPEN
E1-VOLTAGE ACROSS PRIMARY SECONDARY CLOSED
E2-VOLTAGE ACROSS SECONDARY SECONDARY CLOSED

Courtesy RCA

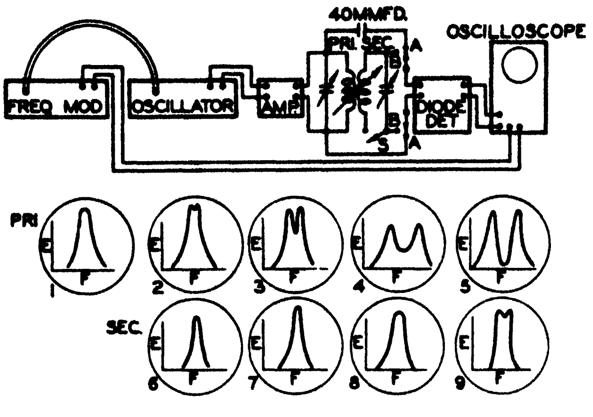
Fig. 19-199.—Arrangement for testing over-coupled circuits.

Coupled Resonant Circuits — Voltages in Over-Coupled Circuits

When the distance is reduced to ⅛ the original value, the primary circuit voltage is equivalent to that obtained with the secondary circuit open. See Fig. 19-199. With the secondary closed, the dip in the primary circuit voltage is very noticeable. Viewing the voltage of the secondary circuit, it is seen that the drop in voltage in the primary more than compensates for the secondary circuit reaction and a dip begins to form in the center of the secondary curve, at the resonant frequency. This is a so-called double-hump curve. When a dip appears in the secondary response curve, the primary and secondary are said to be over-coupled.

Demonstration Setup — Pure Coupled Circuits

The test setup is shown in Fig. 19-200.



Courtesy RCA

Fig. 19-200.—Arrangement for testing pure coupled circuits.

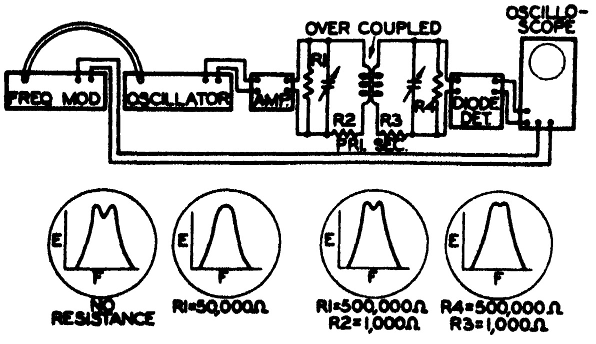
Demonstration No. 9

In this experiment, as in the others where the secondary circuit was open, there is no noticeable reaction on the primary circuit voltage and the curve of primary response remains unchanged with variations in secondary tuning, (open circuit secondary condition). Closing the secondary circuit switches markedly changes conditions in the system.

Patterns *2* to *5* of Fig. 19-200 indicate the effects of increased coupling; as the coupling is made greater, the reaction on the primary is increased, and the dip in the primary response curve becomes greater.

Demonstration No. 10

The rectified voltage resonant curves are shown in parts *6* to *9* inclusive for the various degrees of coupling discussed before.



Courtesy RCA

Fig. 19-201.—Arrangement for testing coupled circuits with resistance.

Demonstration Setup — Coupled Circuits with Resistance

The test setup is shown in Fig. 19-201. Resistance may be added to the coupled circuit system and the effect noted on the cathode-ray-tube screen. Flat-top response curves may be obtained with proper amounts of resistance loading of the tuned circuits. A frequency-modulated signal source is required to carry out the experimental tests. The coupling of the coils is varied to show the effect of different degrees of coupling between primary and secondary circuits. Over-coupling is required to obtain the desired response characteristic in flat-top tuning systems or so-called band-pass tuning arrangements.

Demonstration No. 11

The coupling is increased to secure the over-coupled condition and the double-hump resonance curve is obtained. Placing a 50,000-ohm resistor in parallel with the primary circuit, the resistance loading removes the dip in the response curve but cuts the voltage too much. Installing a larger ohmic value of 500,000 ohms, the circuit loading is reduced; a dip in the response curve occurs, but it is not nearly as pronounced as it was in the absence of the resistive loading.

As an experiment, removing the 500,000-ohm resistor and installing in series with the primary coil a 1,000-ohm resistor, we find that practically the same effect is obtained. With the 500,000-ohm resistance in parallel with the secondary, the dip is considerably smaller than when the same value of resistance was placed across the primary. The same value of resistance has a greater effect when placed across the secondary circuit than when placed across the primary circuit.

Putting 1,000 ohms in series with the secondary, and removing the 500,000-ohm resistance from across the secondary, the same kind of curve is obtained, indicating that either the primary or secondary circuits may be loaded by a series or parallel resistance.

The most desired curve can be obtained, in this series of experiments, by placing an 1,800-ohm resistance in series with the secondary. The desired flat-top curve is thus obtained. This flat-top curve is broader than the optimum coupling curve, which means that better transmission characteristics may be obtained using this type of circuit arrangement.

Panoramic Ultrasonic Analyzer

The Panoramic Ultrasonic Analyzer, Model SB-7, is a scanning heterodyne receiver which *automatically* measures the frequency and amplitude of ultrasonic signals. By means of a stabilized sweeping system, the instrument tunes repetitively six times per second through a 200-kc range in any part of a 10-kc to 300-kc band. As signals are tuned through, they appear on a cathode-ray tube as sharp vertical pips located horizontally in order of frequency. The heights of the pips indicate the relative magnitudes of their corresponding signals. A continuous over-all presentation of the spectrum is obtained on a long-persistence cathode-ray-tube screen. Special control features enable selection and highly detailed examination of narrow bands which may contain signals separated by less than 500 cps.

Direct Reading

The indicating screen provides direct readings of frequency (X-axis) versus signal amplitude (Y-axis). The frequency scale is calibrated linearly through a range of ±100 kc on either side of a center or zero reference frequency, midpoint on the scale. The actual center frequency is indicated on a calibrated tuning dial which permits centering of any portion of the band between 10 kc and 200 kc. When the dial is set for a 200-kc center, signals up to 300 kc can be observed.

Accurate frequency indications are developed by a scanning system which provides driftless displays, constant linearity, stabilized scanning widths, coincidence between tuning dial and screen scale calibrations, and minimum warm-up time.

Both linear and linear-log scales are provided. Selection is made with a front panel switch. The linear scale covers a 10:1 range while the log scale range is 40 db. A fine attenuator control extends the over-all range to over 300:1 or 50 db. Thus signals having amplitude ranges as high as 300:1 can be simultaneously compared.

Full-scale amplitude readings are possible with input voltages between 1 millivolt and 50 volts. Voltages as low as 100 microvolts can be measured when fed directly to the analyzer input. Smaller voltages can be determined by applying them through an external linear preamplifier of known gain. Clear, sharply delineated displays are obtained by utilization of an amber light filter as the calibrated indication screen. Filter action further aids the presentation by reducing the effects of ambient light.

Analysis of Adjacent Signals

Resolution or the ability of separating individual signals for analysis is a function of two factors; the rate of frequency scan and i-f filter bandwidth. Optimum resolution occurs when a definite relationship exists between the two.

Resolution tends to sharpen as the scanning rate and the i-f bandwidth decrease. The scanning rate may be effectively diminished by increasing the scanning period or conversely by decreasing the spectrum width scanned in a given time. The latter method is accomplished in the SB-7 with a logarithmically calibrated scanning width control. A calibrated resolution control varies the i-f selectivity.

As the scanning width is progressively decreased, the deflections immediately surrounding center frequency spread across the screen while those at the outer edges move off scale. Somewhat improved resolution results. Optimum resolution, however, depends upon proper adjustment of the i-f bandwidth with the resolution control.

Thus, the SB-7 is capable of detailed narrow-band analysis of signals which may tend to merge together or even mask one another during maximum scanning width displays. A linearly calibrated center frequency control, permits centering of any group of signals, between 10 kc and 200 kc, for such analysis.

Applications

The SB-7 provides a means for rapid analysis of ultrasonic vibrations in jet engines, electron tubes, ships, missiles or wherever speedy analysis of such vibration is essential. An important feature is that filter bandwidth remains constant as the spectrum is scanned. The instrument may be used for monitoring several telemetering sub-carriers simultaneously.

Ultrasonic sources can be checked against standards for determining frequency and the effects of load changes, component variations, shock, humidity, and thermal changes upon frequency stability. Parasitic oscillations and other spurious oscillations and modulations are easily spotted and identified. Modulation characteristics can be quickly checked by simultaneous comparisons of carrier and sidebands.

The instrument permits continuous monitoring of communications carrier systems for possible off-frequency transmissions, interference due to spillover, spurious modulation, or improper carrier frequency location.

As a frequency selective voltmeter for determining the characteristic of lines and filters, the SB-7 eliminates spurious measurements due to harmonics. As a panoramic indicator, it reduces the time required for characteristic determination.

The instrument can be used for harmonic and cross-modulation investigations since distortion products as low as 50 db down can be evaluated. High resolution capabilities permit direct measurement of cross modulation by comparison of sideband level versus carrier amplitude. Harmonics of the lowest fundamental frequency presented can be completely resolved throughout the entire scanned range.

Fourier analysis of complex ultrasonic waveforms is rapidly provided in graphic visual form for *research* or *instructional* purposes. Presentations are easily interpreted, understood, and remembered.

In general, the instrument may be used in ultrasonic spectrum analysis, harmonic investigations, cross-modulation studies, filter and transmission-line checks, ultrasonic vibration measurements, medical studies, telemetering, and monitoring.

Circuit Description

The instrument consists essentially of an input amplifier, a swept local oscillator, a balanced mixer followed by variable selectivity i-f amplifiers, detector, video amplifier, cathode-ray-tube indicator, associated sweep circuits, and a section for stabilizing the swept oscillator. The block diagram is shown in Fig. 19-202 and the resolution characteristics in Fig. 19-203.

The swept oscillator progressively heterodynes, in order of frequency with those signals at the output of the input amplifier (a phase inverter) to produce a sum frequency which is applied to the i-f section. A highly stable balanced modulator is employed which

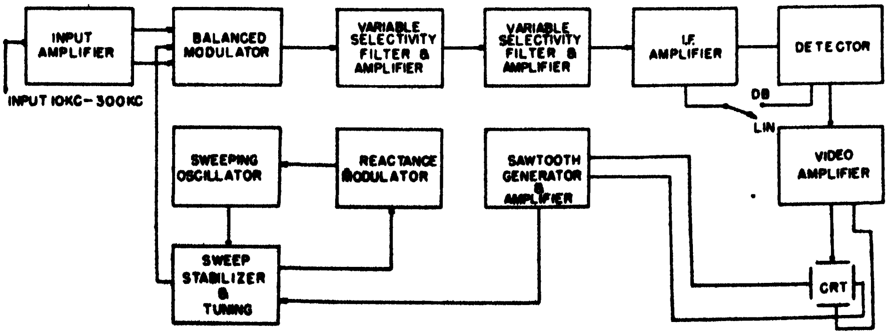
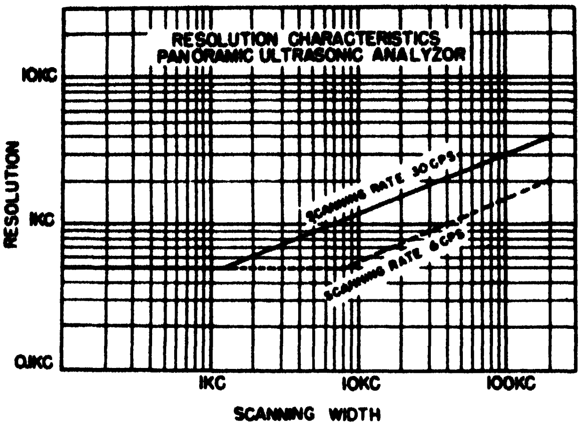

INPUT AMPLIFIER
INPUT 10KC–300KC
BALANCED MODULATOR
VARIABLE SELECTIVITY FILTER & AMPLIFIER
VARIABLE SELECTIVITY FILTER & AMPLIFIER
I.F. AMPLIFIER
DETECTOR
DB
LIN
SWEEPING OSCILLATOR
REACTANCE MODULATOR
SAWTOOTH GENERATOR & AMPLIFIER
VIDEO AMPLIFIER
SWEEP STABILIZER & TUNING
CRT

Fig. 19-202.—Block diagram of the Panoramic ultrasonic analyzer.



Courtesy Panoramic Radio Corp.

Fig. 19-203.—Resolution characteristics of the Panoramic ultrasonic analyzer.

eliminates spurious modulation products and prevents passage of the local oscillator frequency signal into the i-f section. The output voltage of the i.f. which is proportional to the amplitude of the scanned signal, is detected, amplified, and applied to the vertical-deflection plates of the cathode-ray tube. The vertical deflection appears at a definite location along the horizontal axis, according to signal frequency, since a common sawtooth voltage source is used for both the swept oscillator and the horizontal deflection of the cathode-ray-tube beam.

Oscillator sweep is obtained with a sawtooth-modulated reactance-tube circuit. Sweep stabilization is accomplished by feeding the oscillator output to a stabilization circuit which automatically provides correction voltages to the controlling reactor tube. Thus stability of the oscillator center frequency, scanning width, and scanning linearity remain essentially independent of the behavior of oscillator ad reactor circuits. The scanning width is made continuously adjustable by controlling the amplitude of the modulating sawtooth.

The i-f section consists of two variable selectivity stages followed by a tuned stage of fixed selectivity. A ganged control is used to vary the loading on the first two stages.

TRANSDUCERS[27]

Basically, the cathode-ray oscilloscope is a device which plots one electrical voltage as a function of another. From this definition, it would seem logical to infer that the oscilloscope is exclusively a tool of the electrical or electronic engineer. And this assumption would be correct were it not for the *transducer.*

The transducer may be defined as any device which converts a phenomenon into a proportional electrical voltage, so that it may be displayed on the cathode-ray oscilloscope.

The problem of selecting the right transducer for a particular application is an important one, and one which deserves considerable attention, if the best results are to be obtained. Some points to consider in such a selection are:

Method of Energizing

The most desirable transducer is one which is a small, self-contained unit, requiring no bulky, external power supply. Thus transducers such as those employing piezoelectric crystals, which generate the signal voltage without need for a power supply are in common use for many applications. Should a source of power be essential, however, it is advisable to avoid the use of batteries, wet or dry, since these frequently fail or give rise to noise signals in the output at the most inconvenient times.

Transfer Characteristic

An important attribute of a transducer is the relationship between its input and output. In most cases this relationship should be a linear one, a certain change in the input producing a corresponding change in the output, whether this change be an incremental one or a large one. In some cases it is desirable to have a nonlinear relationship between input and output, as, for example, in measuring a phenomenon related to any of the human senses. The relationship between input and output then should usually be exponential, in order to conform with the Weber-Fechner law. In some cases it is desirable to have a logarithmic variation of output with input. An example of this is found where one must measure a very wide range of values, and display them all on the limited scale of the cathode-ray tube.

Sometimes it is desirable to have a trigonometric relationship between input and output. For example one may be interested in displaying the relation between some harmonic and linear motions which are sinusoidally related.

Output Impedance

If, for a given output voltage, one has the choice between low and high impedance in a transducer, low impedance is to be preferred, particularly if the output leads have any considerable length from the transducer to the oscilloscope. High-impedance leads may be used since cathode-ray oscilloscopes, in general, have a sufficiently high input impedance that they will not load the usual transducer sources. But difficulty is encountered in long leads with extraneous electrostatic and magnetic pickup, requiring careful shielding, which usually deteriorates the high-frequency response of the transducer due to the capacitance loading of the shield on the transducer.

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[27] Berkley, Carl, "Transducers, a review of the design and applications of transducers for oscillography," *The Oscillographer,* vol. 12, pp. 9-24, January-March, 1950; the following material is an abridged excerpt of this article.

Amplitude Range

The amplitude range of a transducer should be such that it will not be exceeded by the phenomenon under test. If this is unavoidable, then the output of the transducer should cut off in such a manner that it introduces no spurious signals.

Transducers should also be designed so that off-scale readings will not injure the transducer or cause a change in its zero. This is a typical defect that occurs when the basic element in the transducer is a mechanical one having an elastic limit.

Sensitivity

It is desirable to have the output of the transducer sufficiently high that it may be applied directly to a medium-gain oscilloscope without the necessity of preamplifiers.

Frequency Response

Just as in the design of vacuum-tube amplifiers, a good transducer should have a frequency, phase, and amplitude response sufficient to display the phenomenon under study without distortion.

Mounting Methods

The most convenient and most widely applicable transducer would be in the form of a probe which could be hand-held and inserted into or touched to the object being tested. Probes are convenient for determining the time of occurrence or the waveshape of some devices, but inconvenient when reproducible measurements are to be made. A wide variety of attachment means may be used for transducers.

Other Important Characteristics

The weight of the transducer should be kept to a minimum, since, although a transducer may be designed for measuring the vibration of a massive object such as a beam, it will sooner or later be found desirable to use it for relatively light structures. The size of the transducer should be kept to a minimum for reasons similar to the above. Therefore, the use of subminiature tubes is indicated wherever possible. The shape of the transducer should be such that it can be mounted conveniently with screws, for example, or held by chemical clamps for the sake of versatility. The direction of response of the transducer should be taken into account, since in some cases it is desirable to obtain pickup in a preferred direction to the exclusion of others. Some types of transducers, for example, magnetic coils, have relatively poor directional characteristics in this respect.

A transducer should be designed so as to be capable of continuous operation under the conditions of use without changing its calibration or without deterioration. For example, Rochelle salt crystals and photocells are not capable of long operation under high ambient temperatures and most types of gas-filled photocells are subject to fatigue.

Additional Requirements

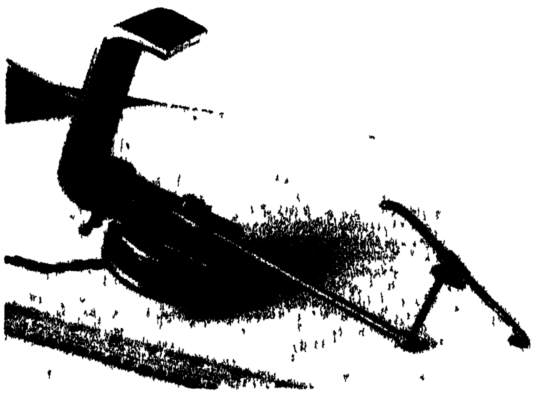
Because of the fact that oscilloscopes are potentially capable of displaying much more information than other types of recorders, the use of an oscilloscope with a particular transducer frequently shows up previously hidden defects in the transducer's output. For example, a microphone used with a sound or film recorder may give perfectly adequate recordings of the peak amplitude of the wave, which is essentially what the ear reacts to. Upon examining the individual waveshapes closely with the oscilloscope, all sorts of distortions and phase shifts in the recorded wave may be seen. Many transducers are not properly damped for the frequencies one is interested in studying. This gives rise to under- or over-shoots in the oscilloscope pattern. One of the best single tests of the frequency-response damping, and phase-shift characteristics of a transducer is the application of a single step function to it. This should be made part of the final test for a transducer before it is put into use. Examination of the output with an oscilloscope will usually show up defects in the shielding of many transducers. The pickup requiring shielding is usually some multiple of the power-line frequency, and not does appear on the average recorder which either does not respond to this frequency or integrates the frequency in order to read only the average, rms or slowly varying value of the phenomenon. A meter, for example, reading the output of an unfiltered power supply will read only the average d-c value; an oscilloscope will show the ripple in the output. The same considerations hold for the signal-to-noise ratio of most transducers. In the case of a low-level transducer, such as a pyrometer thermocouple, a perfectly satisfactory reading is obtained on an ordinary meter, but the oscilloscope shows all the thermal agitation noise which obscures the output reading. In general, it is desirable to choose the transducer such that the signal-to-noise ratio is high enough that gain may be reduced in order that the noise level is limited to a value of about the width of the trace on the cathode-ray tube. It is possible in some cases to use long-persistent screens or photographic methods to integrate the noise from a repetitive pattern.

Types of Transducers

Transducers may be divided into the following classifications:

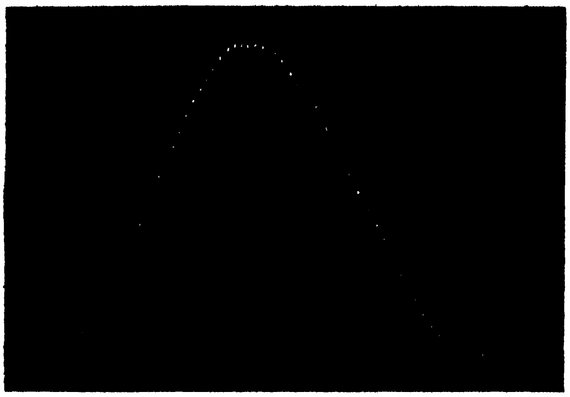
Resistive

The simplest type of transducer is a linear resistor whose resistance is made to change in some manner by the phenomenon under study. The most convenient method of obtaining the change in resistance is to make the resistor a part of a potentiometer or rheostat which is, for example, mechanically coupled to a machine. Fig. 19-204 shows such a typical resistive transducer, designed for measuring the aperture or "shed," between certain crossed threads on a loom. The right-angled foot with upturned edges rests upon a group of threads and reads their average position by varying the resistance of the potentiometer. It is usually convenient to use direct current for energizing the potentiometer, although alternating current may be used if only an a-c amplifier is available, or if no source of d-c is available. The line-voltage output available on standard oscilloscopes is convenient for this purpose. Fig. 19-205 shows a typical shed aperture versus time curve obtained from this transducer. Angular timing markers have been superimposed on the pattern by means of beam modulation.



Courtesy Du Mont Labs.

Fig. 19-204.—Typical resistive transducer for measuring the shed aperture on a loom.



• Courtesy Du Mont Labs

Fig. 19-205.—Oscillogram of shed-aperture versus time, obtained by use of transducer of Fig. 19-204.

Potentiometers suffer from the defect that, at high frequencies, the potentiometer arm tends to bounce from wire to wire of the card, giving rise to discontinuities in the output. This can be eliminated in some cases by a shunting capacitor, but this tends to cause phase shifts in the output. Carbon potentiometers may also be used, but these tend to wear out rapidly with continuous use. A better solution is the use of single slide wires for the potentiometer. Such slide wires are capable of being made as long as desired or arbitrarily shaped to fit a particular machine disposition.

Resistive transducers need not be wires. For example, a humidity transducer has been made by using a strip of cotton fibers impregnated with a saline solution, which either effloresces or deliquesces, depending upon ambient humidity. Such a sensing element has been used in meteorological balloons to determine cloud heights. Nonlinear resistors also form suitable transducer elements in many cases. The strain gage is an example of such a transducer. The cross section of the gage wire, which is closely bonded to an element whose strain is being measured, changes with extension or compression of the measured member, resulting in a variation in electrical output. Such strain gages have been either bonded directly to objects being measured, such as girders or incorporated upon extensible gage elements. One defect of resistance strain gages is that the percentage of change in resistance with elongation is small and this requires the use of a balanced bridge to avoid stray pickup, and the use of carrier signals to obtain high amplification with tuned amplifiers. D.c. may be used where careful attention is paid to the shielding and where large signal amplitudes are expected. The frequency range of this type of transducer is normally of the order of a few kc. Recently developed types of conductive rubber (U. S. Rubber Company) show some promise in the design of nonlinear resistive transducers capable of wider ranges of elongation. When using a bridge type of strain gage carrier modulator, it is desirable to know whether the unbalanced output indicates compression or extension of the gage. This may be done conveniently by marking a particular portion of the carrier phase with a blanking dot. The carrier then has a dark or bright line on the top or bottom half of the wave for compression or extension.

Coil springs which unwind progressively, one coil at a time, have been studied by the National Bureau of Standards, and appear to be capable of use as nonlinear resistors over a wide range of amplitudes and resistances.

Varistors are another type of resistive element which change in resistance, depending upon the impressed voltage or current. These are useful as current or voltage transducers over a wide range of magnitudes in order to display over-all values on the face of the tube.

Thermistors are similar variable resistances which change value with temperature. These are, therefore, capable of use as sensing elements for changes in temperature. They also are more sensitive in bridge circuits. Various effects other than temperature may be measured by means of the change they produce in the temperature of some object. For use with an oscilloscope, a thermistor to measure temperature may merely be connected between the input of the oscilloscope and a source of supply voltage applied through a resistor whose value is made as high as practicable, consistent with the amplification of the oscilloscope and the pickup encountered. The supply voltage may be either d.c., a carrier frequency, or the 60-cps test signal supplied by the oscilloscope. Thermistors may be obtained with time constants due to thermal lag of as low as 1/10 second.

Switches are convenient transducers for many oscilloscopic tests. Vibrating switches are sometimes useful as choppers for a d-c signal, in order to enable its amplification through an a-c coupled amplifier. This use is in general limited to very low frequencies, however, where the advantage of the oscilloscope is not so apparent. At high frequencies, the switch cannot be made to operate efficiently and electronic switches must be used. Switches are also useful to produce markers upon the trace indicating the instant of operation of some portion of a machine. By means of differentiating and integrating circuits, the pulses from various switches may be given different waveshapes or characters in order to enable a large number of phenomena to be displayed upon a single time base.

Electrostatic Transducers

A variation in capacitance may also be used to produce an electrical signal. From the formula $Q = CE$, either Q, C, or E may be changed to produce an indication on the oscilloscope. One of the first transducers used was the condenser microphone. This consists of a thin diaphragm vibrated by the sound which has a capacitance to the grid of a vacuum-tube amplifier. With a charge on the diaphragm, its motion induces a resultant motion of the electrons to and from the grid of the tube giving rise to the voice modulated signal. This effect increases in sensitivity as the spacing is made smaller and increases with the applied voltage. This effect is, therefore, most suitable for small spacings. Vibrating reed capacitors, in conjunction with a-c operated amplifiers, have also been used to amplify d-c signals which are impressed upon the capacitance. Capacitors have been used as transducers in oscillators whose frequency is varied by the capacitance. The oscillator frequency may be varied by the space between the plates measuring a distance or by the change in the dielectric constant or dipole moment of the substance between the electrodes.

Capacitors cut in special orientations are also used as transducers in sweep frequency generators, in which the rotation of the capacitor or displacement is made a function of the desired frequency. Specially cut capacitors of this type have also been used as continuously variable phase shifters for cathode-ray-tube presentations. Such transducers can be made to have a phase shift which is accurately proportional to the angle through which a shaft is turned.

Another type of transducer which may be classified here is the mechano-electronic transducer. This produces a variable output depending upon the position of a magnetically variable plate in a triode tube. This transducer is quite sensitive to small angular or linear displacement but suffers from the defect mentioned above, that when its elastic limit is exceeded, which can be done quite readily, it does not return to zero. The frequency response of this unit is of the order of 12 kc. Many phenomena can be converted to a mechanical displacement and, therefore, measured with this transducer.

Electromagnetic Transducers

A coil moving in a magnetic field can obviously be used as a source of electrical voltage. This voltage is usually proportional to the velocity or rate of change of the motion and, therefore, requires integration in order to produce a true indication of displacement.

Another type of magnetic transducer depending upon the motion of a moving permanent magnetic field and a stationary coil is a two-phase a-c generator used to transduce rotational speed. The a-c output of this pickup is accurately proportional to the speed of its center shaft. The generator has two phases which are 90° apart and is, therefore, useful in generating a circle on a cathode-ray tube which rotates at the same speed as a machine to which it is coupled, giving an automatically synchronized circular sweep.

There are many types of magnetic transducers. A simple coil can detect the motion of most ferrous objects due to their residual magnetism. The characteristics of a ferromagnetic material are not usually altered for industrial purposes if it is magnetized in order to enable electrical pickup of its motion. For example, a steel projectile may be magnetized and its time of passage through two coils generate two marker voltages which enable an accurate determination of the shell's muzzle velocity. In a similar manner to that for capacitor transducers, a variable coil may be made the frequency-determining element in an oscillator. If desired, the oscillator may also be varied over a range of frequencies and the response of the part under test noted to be compared with a standard material. In order to obtain a linear response from a variable-coil-determined oscillator, the output is generally detected using a detector operating on the linear portion of the response curve of the output transformer.

Most types of transformers can be used as electromagnetic transducers in one fashion or another. A variometer whose output is proportional to some function of its angle may be used to detect the motion of some object or machine. The common selsyns may be used as such transducers with the restriction that the output is a sinusoidal function of the angle through which the shaft is turned. A particularly useful type of transformer transducer is the differential transformer. These transformers are available in a wide variety of shapes and capacities. They consist of a single primary and two symmetrically wound, opposed secondaries. These transducers may be used, for example, to detect slight changes in some material capable of producing an inductive effect. If the material is uniform and the transformer balanced for zero output, then the output will remain zero unless a change in the characteristics of a portion of the material running through the transformer occurs. This particular transducer will work just as well on magnetic and nonmagnetic metallic materials. The nonmagnetic materials act as shorted turns, which have identical effects on both ends of the transformer. This type of transducer can also be used as a position transducer in which a slug of material is brought closer to one coil than another. It can also be used to detect changes in the thickness of materials by placing one end of the transducer against the material and having it run past.

Saturable reactors of various types can be used as transducers for small currents or changes in reluctance to produce relatively large voltage changes which are capable of display on oscilloscopes. Such saturable reactors generally have a rather poor frequency response for oscilloscopic purposes, but have the advantages of great ruggedness for industrial applications and the fact that they do not require any vacuum tubes for their operation.

Magnetic field-strength detectors are frequently desired to measure the strength of a steady or d-c magnetic field in space. This is difficult to do in a continuous fashion. One simple method is to rotate a coil in the field and measure on an oscilloscope the peak value of the induced voltage. The value of the field can then be calculated from an observation of the variation of the output with the angle of the coil on the oscilloscope. The cathode-ray tube itself can be used as a d-c magnetic probe but is usually relatively large for this purpose.

A small magnetic transducer accomplishing the desired variation of output with magnetic field intensity is the magnetically controlled diode. This is a simple diode in which the variation of plate current with ambient field intensity is roughly linear over a certain range. This can be used, for example, as a position transducer by having the position of a small bar or horseshoe magnet in the vicinity change its position with respect to the tube. It may be used as a pressure transducer too, for example, by attaching a small magnet to the end of a Bourdon tube placed near the diode.

Piezoelectric Transducers

The piezoelectric crystal is one of the earliest forms of transducers. Either natural or synthetic crystals may be used. The natural crystals have the advantages of very low leakage which with an electrometer input enables the measuprements of very slowly varying pressures. These natural crystals, however, have the disadvantage of very low sensitivity. The natural crystals are to be preferred where very hard usage is to be encountered, or where high temperatures or sudden shocks occur. The synthetic crystals have the advantage of very high output for a given displacement, pressure, or strain but these are usually temperature sensitive and are not as capable of withstanding high applied forces without fracture. The natural crystals, in general, have a higher frequency response, while synthetic crystals may be made quite large and, therefore, can be used, for example, as frequency-determining elements in relatively low-frequency oscillators.

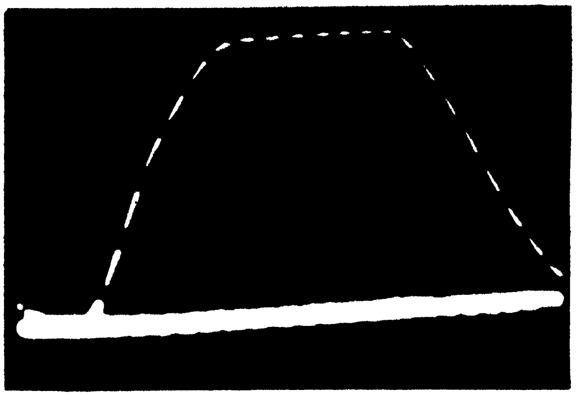
A new class of synthetic materials such as the barium titanates has recently become commercially available, and has been very widely used in phonograph pickup transducers. This material has the advantage for transducing, that it may be readily formed into special shapes which make it adaptable, for example, to the design of hemispheric-shaped sound transducers for concentrating their attention upon a small area at the focus of the sphere. These synthetic materials are also relatively fragile and temperature-sensitive.

Thermoelectric Transducers

The thermoelectric effect is another capable of being used for transducers. The thermocouple or the optical pyrometer, in which the image of the furnace is focused upon the thermocouple are two examples. The difficulty with most thermocouples is their heat lag, which rules them out for any high-frequency heat changes. A photocell pyrometer is usually to be preferred, if the temperature is such that the photocell effect can be used.

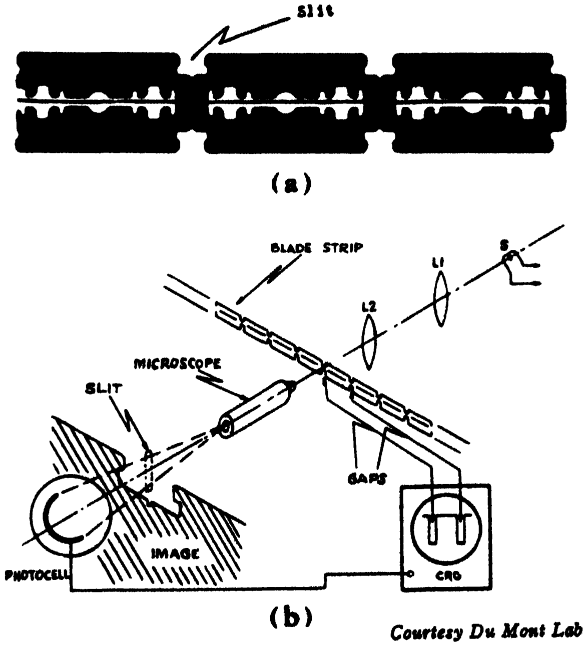
Photocell Transducers

The photocell is probably the most versatile transducer known. A photon causes the least loading upon the source, compared to any other transducer effect. Photocells of the barrier layer type, such as the Weston Photronic Cell, have the disadvantage of low voltage sensitivity and poor frequency response, but have the advantage of very good stability for industrial applications. Resistive photocells, such as the selenium type, which are too unstable for direct calibration, and the gas or vacuum field photocells which are subject to the usual gas oscillations and fatigue are to be avoided in general. Vacuum photocells are most suitable for oscilloscopic use. They may be obtained in a range of sensitivities from the small probe-type 1P42 units, to the high-sensitivity multiplier photocells. In designing a photocell transducer for oscilloscopic use, account should be taken of the electrode capacitance of the photocells, since the high-value load resistors used with the cells may be shunted appreciably, even at fairly low frequencies. Care should also be taken in designing the light sources for photocells so that even illumination for the cathodes is provided to avoid variations in signal resulting from variations in sensitivity of portions of the cathode surface. The obvious application for a photocell on the oscilloscope is to show light output versus time curves. A typical curve of shutter opening versus time is shown in Fig. 19-206.



Courtesy Du Mont Labs.

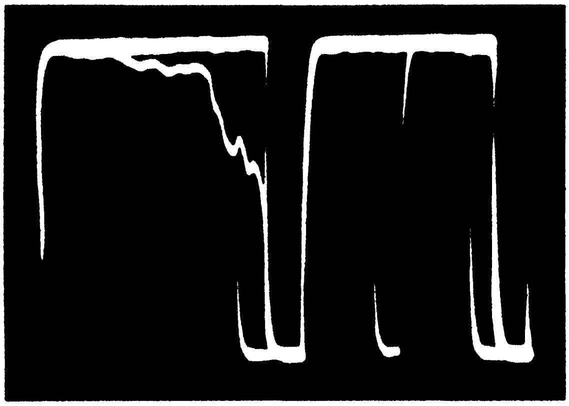
Fig. 19-206.—Oscillogram indicating camera shutter opening versus time, obtained by use of a photoelectric transducer.



Courtesy Du Mont Labs.

Fig. 19-207.—(A) Razor-blade strip; (B) optical arrangement for scanning razor-blade strip.

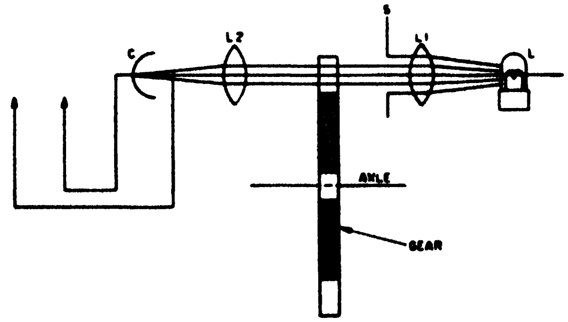
It is frequently convenient to scan an object, outline, or surface, using photocells. Fig. 19-207 illustrates such an instance. Here, the outlines of razor blades which are prepared in a continuous strip are examined by moving them past the photocell. Fig. 19-207B shows the optical setup for scanning the razor-blade edges using a microscope to magnify the images and a slit for examining a small portion of the edge at any instant.



Courtesy Du Mont Labs.

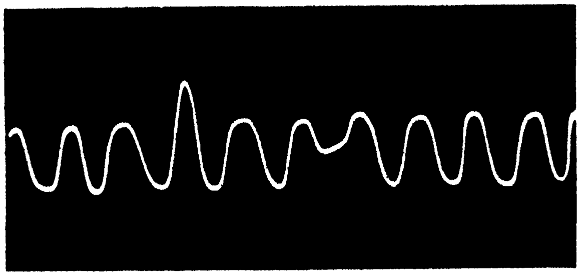
Fig. 19-208.—Oscillogram showing razor-blade edges, some of which are nicked.

Fig. 19-208 shows a typical oscillogram made from razor-blade edges, revealing a number of nicks and other defects. Fig. 19-209 shows another photocell scanning method for testing a rotating gear, in which the optical system is so arranged that the output of the photocell is proportional to the height of each gear tooth above the base circle of the gear. Fig. 19-210 shows a typical oscillogram showing the waveshape from a number of teeth, two of which are defective. With a more rapid sweep, synchronized with the repetition rate of the gear teeth, we obtain the patterns of Fig. 19-211, enabling the comparison for congruence of all the teeth.



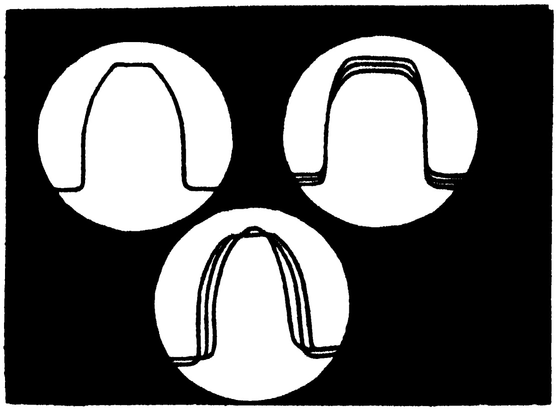
Courtesy Du Mont Labs.

Fig. 19-209.—Optical setup for inspecting gear teeth.



Courtesy Du Mont Labs.

Fig 19-210.—Oscillogram made from setup of Fig. 19-209, showing that two of the gear teeth are defective.



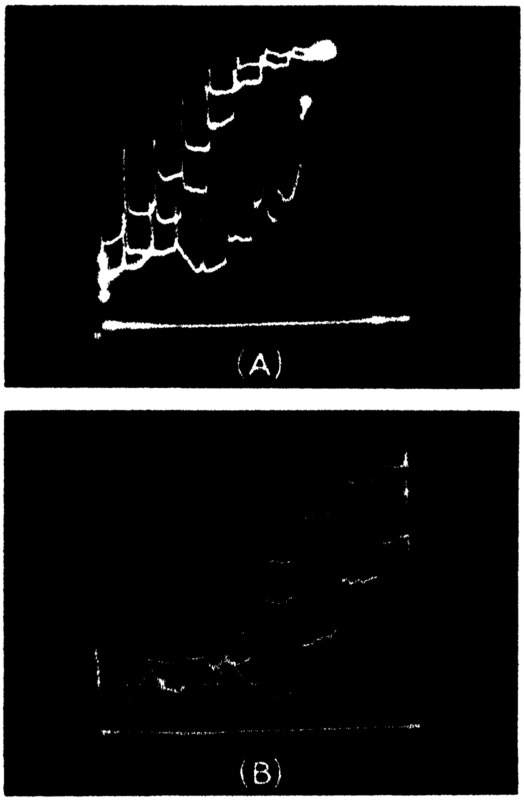
Courtesy Du Mont Labs

Fig 19-211.—More rapid sweep enables comparison of patterns for congruence of teeth. Top left, perfect gear; top right, eccentric shaft; below, irregular teeth.

Courtesy Du Mont Labs.

Fig. 19-212.—Experimental setup for tracing H and D curves.

In photography it is desirable to draw automatically H and D curves, illustrating the variation in density of sensitometric strips. It is desirable also to obtain this information indicating how this density varies during development. Fig. 19-212 shows an experimental setup for H and D curve tracing of a circular strip rotated together with a mechanically linked potentiometer used as an angle transducer. A photocell operating on non-actinic light is used to pick up the reflectance of the paper sensitometric strip. Fig. 19-213 shows two sets of typical H and D curves obtained during development.

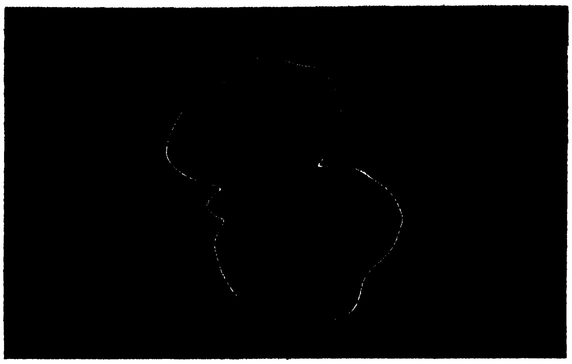


Courtesy Du Mont Labs.

Fig. 19-213.—Two sets of typical H and D curves obtained during paper development.

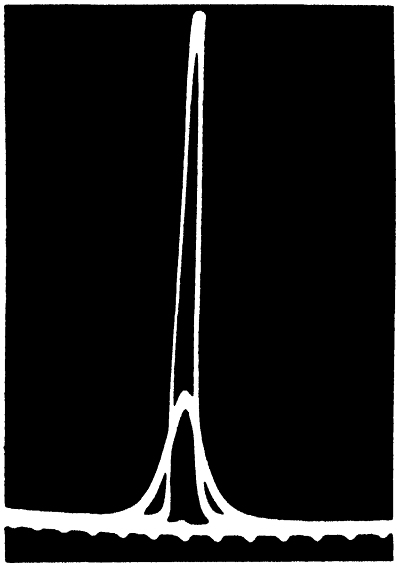
An oscilloscope and photocell offer a convenient combination for plotting the light-distribution curves of various surfaces. Fig. 19-214 shows such a typical light-distribution curve in polar co-ordinates made on a Du Mont Type 275-A Polar-Co-ordinate Indicator.

A convenient method of scanning surfaces is a vibrating mirror. This mirror is mounted on a tuned resonant reed, oscillating at the power-line frequency. Fig. 19-215 shows typical light-distribution curves made in this manner, from glossy and matte surfaces.



Courtesy Du Mont Labs.

Fig. 19-214.—Typical polar-co-ordinate light distribution pattern.

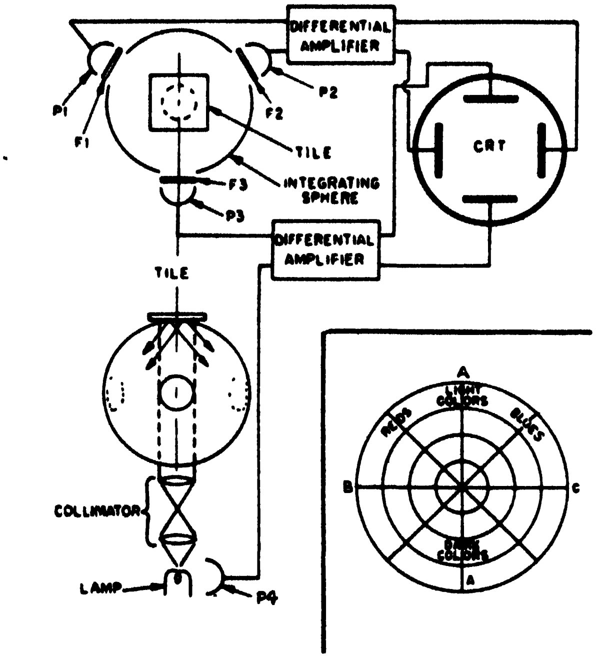


Courtesy Du Mont Labs.

Fig. 19-215.—Two superimposed light distribution curves, one from a matte surface, the other from a glossy surface.

The X-Y presentation on a cathode-ray oscilloscope enables the simultaneous presentation of two values, which is of advantage in a particular photocell application which presents both the brightness and color of a surface as indicated by the X-Y position of the spot on the cathode-ray tube. This is similar to the I. C. I. presentation of color co-ordinates, except that the color primaries have been chosen arbitrarily. Fig. 19-216 shows schematically the type of connections required for the three photocells, two of which read color, the third one luminosity.

The above-mentioned applications which have necessarily been considered briefly, show a wide range of problems to which photocell transducers have been applied. These illustrations represent only a small fraction of the potential ones.



Courtesy Du Mont Labs.

Fig. 19-216.—Schematic diagram of connections required for three photocells, two reading color, and one reading luminosity.

Choice of Transducers

As should now be obvious, for any particular application, one has the choice of a wide variety of transducers or a wide variety of electrical or related effects to choose from in measuring any phenomenon. In choosing the best transducer, the points mentioned at the beginning of this article: method of energizing, transfer characteristic, output impedance, amplitude range, sensitivity, frequency response, mounting method, size, weight, shape, durability, and signal-to-noise ratio should be considered in order to obtain optimum results. Let us apply some of these criteria to the measurement of typical phenomena, which arise similarly in many related fields, in order to determine which type of transducer would be most suitable.

Displacement

Displacement may be measured by means of a slide wire, or a differential transformer; or a seismic suspension might be used with a piezocrystal. However, if the above methods load the source excessively, it may be desirable to use a photocell and optical system.

Vibration

Similarly to displacement, vibration may be measured photoelectrically, piezoelectrically, magnetoelectrically or electrostatically. The most convenient method will depend upon the conditions of the particular problem.

Light

The only suitable method for measuring light for oscilloscopic purposes is by means of a photocell.

Temperature

Temperature could be measured by means of the thermoelectric effect, a thermistor, or a photocell. Again the photocell is to be preferred if rapid changes are expected or if the temperature is such as to radiate detectable wavelengths.

Pressure

Pressure may be measured by means of strain gauges, by the gas-tube transducer, or by means of a photocell and diaphragm-actuated mirror.

Humidity

Humidity, as indicated above, can be measured either photoelectrically by means of a mirror upon which condensation takes place, or by means of a variable resistance salted fiber.

Vapor Pressure

Vapor pressure might be measured by means of the Toricelli vacuum displacement, either with a photocell or by having the mercury column control and oscillator, or the output of a differential transformer.

Sound

Sound may be measured by a diaphragm actuating a magnetic, electrostatic, or photoelectric transducer.

Strain

Strain may be measured either resistively, capacitively, or magnetically.

Torque

Torque may be measured magnetically by the Cox torque transducer which consists of two spring-coupled rotating masses, magnetically coupled. Variation of torque causes a relative motion of the masses which gives rise to an electrical signal. Torque may also be measured photoelectrically. The bar or shaft whose torque is to be measured has connected to it two crossed polaroids, with a light beam to a photocell passing through them. Torque in the bar causes a twist of the polaroids which gives rise to sinusoidal variation in light intensity with torque.

CHAPTER 20

COMPLEX WAVEFORM PATTERNS

In the design and development of electronic equipment, the voltages and currents encountered will, in many cases, be nonsinusoidal. In almost all cases, the oscilloscope will be used for visual observation of the waveform of these voltages and currents. The nonsinusoidal or complex waveform may represent signals that are not desired, as in the case of distortion in an amplifier; or it may be the normal type of signal for a particular circuit as the waveform of a musical sound. In either case, the nonsinusoidal waveform can be represented by the sum of a number of pure sine waves, consisting of a fundamental and harmonic frequencies.

The harmonic content of a complex wave can be determined by several different methods, among which are the use of the Fourier series and harmonic wave analyzers, or comparison with a complex wave which has been synthesized and whose harmonic content is, therefore, known. The last method is the one that will form the basis of this chapter.

The use of the Fourier series to determine the harmonic content of a complex wave is usually rather laborious except in the case of waveforms that are regular in shape, as square-wave, sawtooth, triangular and similar waveforms, when the wave can be expressed by familiar mathematical functions. These waveforms have all been analyzed by the use of the Fourier series, the results of which can be found in any text or handbook. For the analysis of irregular waves such as the distorted output of an amplifier, graphical methods can be used to obtain the coefficients of the Fourier Series. In either case, the amplitudes and phase of the harmonics present in the wave will be obtained, but the time required for the necessary calculations will impose limitations upon the Fourier method of wave analysis.

The use of a harmonic wave analyzer has the advantage of greatly reducing the time required to obtain the amplitudes of the harmonic components of a complex wave, as compared to the Fourier method. There are two general types of harmonic wave analyzers; they are the mechanical and the electrical type. The mechanical type requires a drawing of the complex wave, so that a pointer of the analyzer can trace the wave. The electrical type operates directly from the signal itself which is fed into the analyzer. The mechanical analyzer is not as fast as the electrical type but will determine both the amplitude and phase of the harmonics, whereas the electrical analyzer will determine only the amplitude of the harmonics. Another difference between the two is that the mechanical analyzer is not limited by the frequency of the wave as long as a drawing of the waveform can be obtained. The electrical analyzer, on the other hand, is limited to a frequency range that the necessary internal circuits will handle, which for present-day electrical harmonic wave analyzers is the a-f range. There is one feature that both the mechanical and electrical type of analyzers have in common; they are both fairly expensive and not always readily available.

In a great many cases where harmonic wave analysis is required, the time and/or cost involved of the two methods mentioned above will be prohibitive. It is mainly for these cases that the waveform patterns in this chapter have been prepared, though, undoubtedly, many other uses will be found for these patterns.

The synthesized waveform patterns have the advantage of being quick and easy to use and do not require any special equipment other than an oscilloscope on which the signal is observed. They show the effect of both amplitude and phase changes at a glance. Intervals between amplitude and phase changes are small enough so that intermediate values may be readily obtained by interpolation. The number of different harmonic combinations shown have been limited to those which appear to be most useful and likely to be encountered in actual practice. It can be seen that, theoretically, an infinite number of patterns are possible, since both the number of harmonics possible and the number of combinations of various amplitudes and phases are both infinite. Actually, however, the shape of any complex wave is determined mainly by a few low-order harmonics, and the number of possible phase-amplitude combinations can be reduced to reasonable numbers by using discrete intervals.

It may be interesting to note here that all the waveform patterns shown in this Chapter are actual photographs of oscilloscope patterns. This may not be apparent, since they were printed in reverse colors; that is, black on a white background instead of white on a dark background as the pattern would normally appear on the screen of the cathode-ray tube. This was done to facilitate make-up of this particular part of the book, but does not in any way interfere with the usefulness of the pattern. The waveforms were obtained by generating the fundamental and harmonic frequencies with the necessary phase and amplitude relationships, mixing them and applying the resultant signal to the input of an oscilloscope. The pattern on the oscilloscope was then photographed for each desired combination. The apparatus used to generate these complex waves will be described a little further along.

Use of Waveform Patterns

The complex waveform patterns displayed on the following pages consist of three types of combinations. These are: fundamental and one harmonic, fundamental and two harmonics, and two harmonics without the fundamental. The various combinations of harmonics within these groups and the pages that they start on are as follows:

	Page
Fundamental and one harmonic	
Fund. plus 2nd harmonic	751
Fund. plus 3rd harmonic	753
Fund. plus 4th harmonic	755
Fund. plus 5th harmonic	757
Fund. plus 6th harmonic	759
Fund. plus 7th harmonic	761
Fund. plus 8th harmonic	763
Fundamental plus two harmonics	
Fund. plus 2nd and 3rd harmonics	765
Fund. plus 2nd and 4th harmonics	785
Fund. plus 3rd and 5th harmonics	805
Two harmonics, no fundamental	
2nd plus 3rd harmonic	825
2nd plus 5th harmonic	826
3rd plus 5th harmonic	827
4th plus 5th harmonic	828
5th plus 6th harmonic	829

In each combination of frequencies, all possible combinations of amplitude and phase are shown. Five amplitude intervals are used in every case. They are 5, 15, 30, 60, and 100 per cent. Eight phase intervals are used for the combinations of fundamental and one harmonic. They are 0°, +45°, +90°, +135°, 180°, −135°, −90°, and −45°. For the combinations of fundamental and two harmonics, and two harmonics without fundamental, four phase intervals are used. These are 0°, +90°, 180°, and −90°.

Where the waveform contains the fundamental and either one or two harmonics, the fundamental is used as the reference signal whose amplitude will be 100 per cent and whose phase is 0° for all combinations. The amplitudes and phase angles of the harmonics will vary at the intervals stated with reference to the fundamental. Where the waveform contains only two harmonics without the fundamental, the lower harmonic is used as a reference and stays constant at 100 per cent amplitude and 0° phase.

The waveform patterns have been arranged on the following pages so that any desired pattern can be found easily. The caption on the top of each page shows the harmonics that all the waveforms on that page contain and also the extent of variation of phase and amplitude of each harmonic. It will be noticed that on any one page, amplitude and phase variations take place for only the higher-order harmonic where two harmonics are used, with the phase and amplitude of the lower-order harmonic staying constant for the entire page, but changing from page to page.

The values given for the variation in phase and amplitude are inclusive. Where the amplitude variation is given as 5 to 100 per cent, this indicates that the page includes all the amplitude intervals, which are 5, 15, 30, 60, and 100 per cent. The phase varies in a positive direction across the page and also from page to page. Therefore, where the phase variation is given as 0° to −90°, it will include all phase intervals between 0° and 360° in a positive direction, which are 0°, +90°, 180°, and −90°. The waveform for 360°, will be the same as for 0°. Note that negative angles are used for values between 180° and 360°. If desired, positive angles can be substituted, such as 270° for −90°, etc.

Another feature of this arrangement is that the amplitude of the higher-order harmonic increases from 5 per cent at the top of a column to 100 per cent at the bottom, while the phase for the column remains constant. Similarly, the phase angle increases from left to right on each row by either 45° or 90° for each pattern, while the amplitude remains constant for the row of patterns. This is helpful when determining the effect of a change in either amplitude or phase. Also, by observing the manner in which the shape of the waveform progresses for changes in either phase or amplitude, it will be possible to determine the shape of the waveform for values of phase and amplitude other than those shown.

The complete expression for any of the complex waveforms can be written as

$$e = A_0 + A1 \sin(\omega t + \theta) + A2 \sin(2\omega t + \theta 2)$$
$$+ A3 \sin(3\,\omega t + \theta 3) + \ldots.$$
$$+ A3 \sin(3\,\omega t + \theta 3) + \ldots.$$

where e = instantaneous value of wave
$\omega = 2\pi f$ (f being the fundamental frequency)
t = time
A_0 = amplitude of d-c component
$A1, A2, A3 \ldots$ = amplitude of fundamental and corresponding harmonics
$\theta 1, \theta 2, \theta 3 \ldots.$ = phase angle of fundamental and corresponding harmonics.

The d-c component A_0 is not given for the waveforms, as it does not affect the shape of the waveform and the measurement of this value does not present any problem. $A1$ and $\theta 1$ for the waveforms shown will always be 1 (or 100 per cent) and 0°, respectively. If the waveform is to be determined for a given expression where $A1$ and/or $\theta 1$ are not these values, it will be necessary to rearrange the expression so that these values are obtained. This is easily done by dividing through by $A1$ and by shifting the phase by an angle equal to $-\theta 1$. In doing this, it should be remembered that $\theta 2$ will change twice as much as $\theta 1$, $\theta 3$ will change 3 times as much, etc.

In the waveforms that contain two harmonics without a fundamental, $A1$ becomes zero and the lower-order harmonic is used as the reference. In this case, it may be necessary to rearrange the expression so that the lower-order harmonic has an amplitude of 1 and a phase of 0°. The procedure is similar to the case of the fundamental, except that the amount of shift for the higher-order harmonic is equal to the phase shift for the lower-order harmonic times the ratio of frequencies. For example, if the waveform contains only 2nd and 3rd harmonics, and the 2nd has to be shifted 30° to make its phase equal to zero, then the 3rd harmonic will have to be shifted 45° in the same direction.

Complex-Wave Generator

The basic requirements for a complex-wave generator are synchronization of the fundamental and harmonics and independent control of the phase and amplitude of each harmonic component of the complex wave. The equipment used to generate the waveforms in this chapter is shown in the block diagram of Fig. 20-1. The diagram shows the complex-wave generator as having 8 channels, which indicates that it will generate signals that contain a fundamental and harmonics up to the eighth. However, the same type of equipment has been used to supply harmonics up to the 12th by adding four more channels to the existing eight.

The operation of the complex-wave generator is apparent from the block diagram. Starting at the extreme left, we have a 1,000-cps frequency standard which is used to synchronize the multivibrator of channel 1 to exactly 500 cps. The output is a square wave which is fed both to a tuned amplifier and to a differentiator and clipper. The differentiator produces pulses from the square wave fed to it, but these pulses are positive and negative, so a clipper is used to remove the positive pulses. The remaining negative pulses are fed to the multivibrators of all the remaining channels and also to an external terminal to supply a sync signal for an oscilloscope. This 500-cps sync pulse synchronizes the multivibrators of each channel from the 2nd to 8th at its proper operating frequency. The 2nd channel will sync at 1,000 cps, the 3rd at 1,500 cps, etc.

From this point on, all the channels are alike except for the operating frequency, so it will be sufficient to

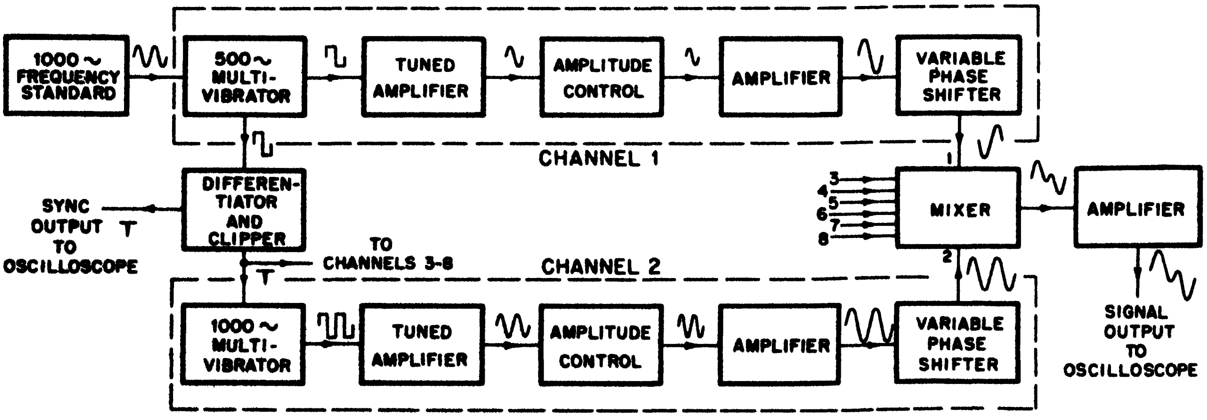


Fig. 20-1.—Block diagram of the complex-wave generator used for taking the waveforms shown in this chapter.

discuss only one. Returning to the 500-cps multivibrator in channel 1, the square-wave output is fed to a tuned amplifier which, in this channel, is tuned to 500 cps. The output of this amplifier will be a 500-cps sine wave, which is fed to another amplifier, which is not tuned. The gain control in the input circuit of this 2nd amplifier is the amplitude control for the channel. The signal is then fed to a capacitance-type phase shifter which can vary the phase from 0° through 360°. After the phase shifter, the signal is fed to a mixer, together with the outputs of all the other channels. The composite signal is then further amplified and then fed to the vertical input of an oscilloscope.

The horizontal sweep of the oscilloscope is synchronized with the sync signal supplied by the complex-wave generator so that several cycles of the complex waveforms are observed. For the waveforms shown, three cycles of the fundamental frequency were shown. Each waveform desired was then set up by adjusting the phase and amplitude controls of each channel to their proper setting. For harmonics not being used, the amplitude control was set at zero. The resulting waveform was then photographed from the pattern displayed on the cathode-ray tube. This procedure was repeated for every one of the complex waveforms shown.

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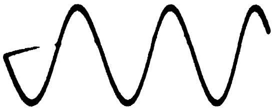
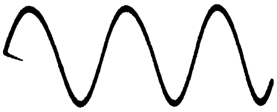
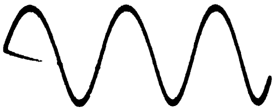
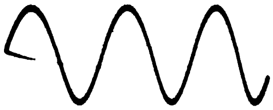
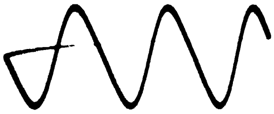
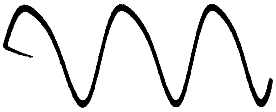
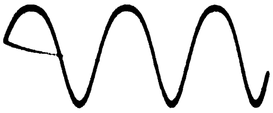
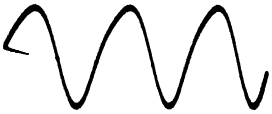
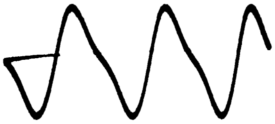
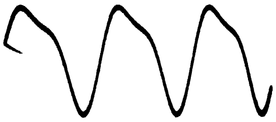
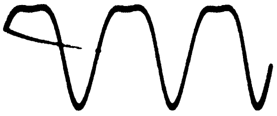
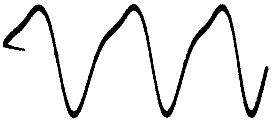
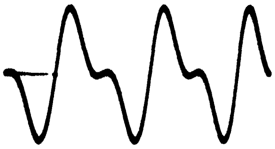
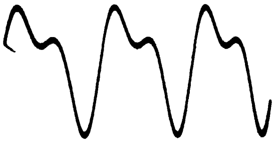
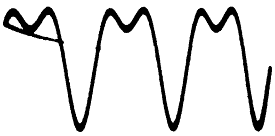
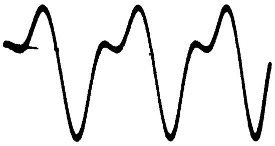
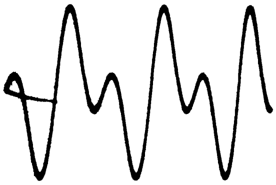
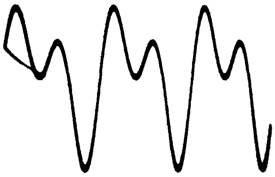
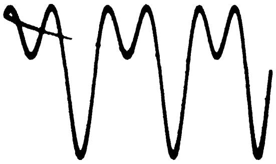
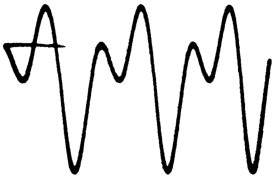
 A=5%	 A=5%	 A=5%	 A=5%
 A=15%	 A=15%	 A=15%	 A=15%
 A=30%	 A=30%	 A=30%	 A=30%
 A=60%	 A=60%	 A=60%	 A=60%
 A=100%	 A=100%	 A=100%	 A=100%
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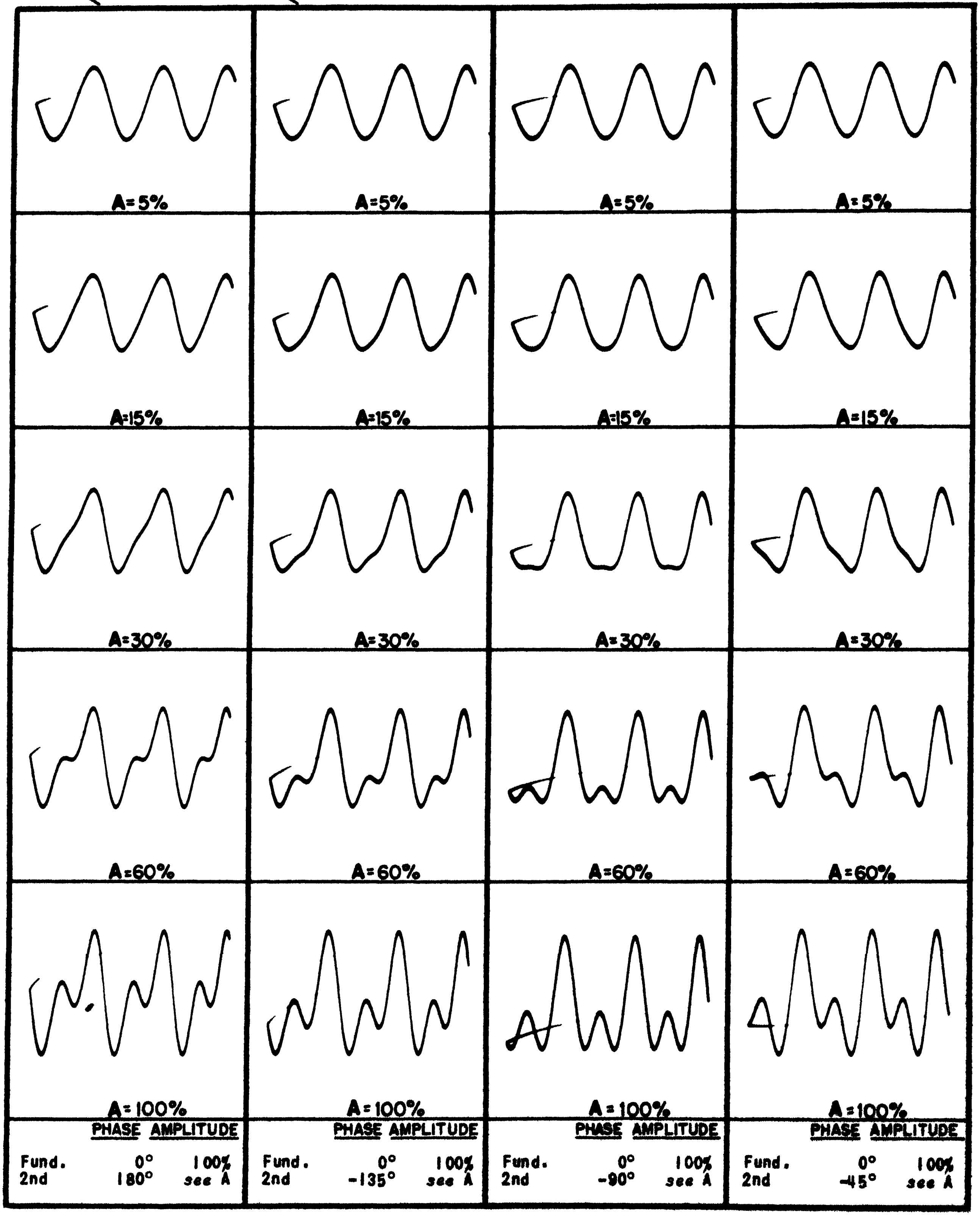

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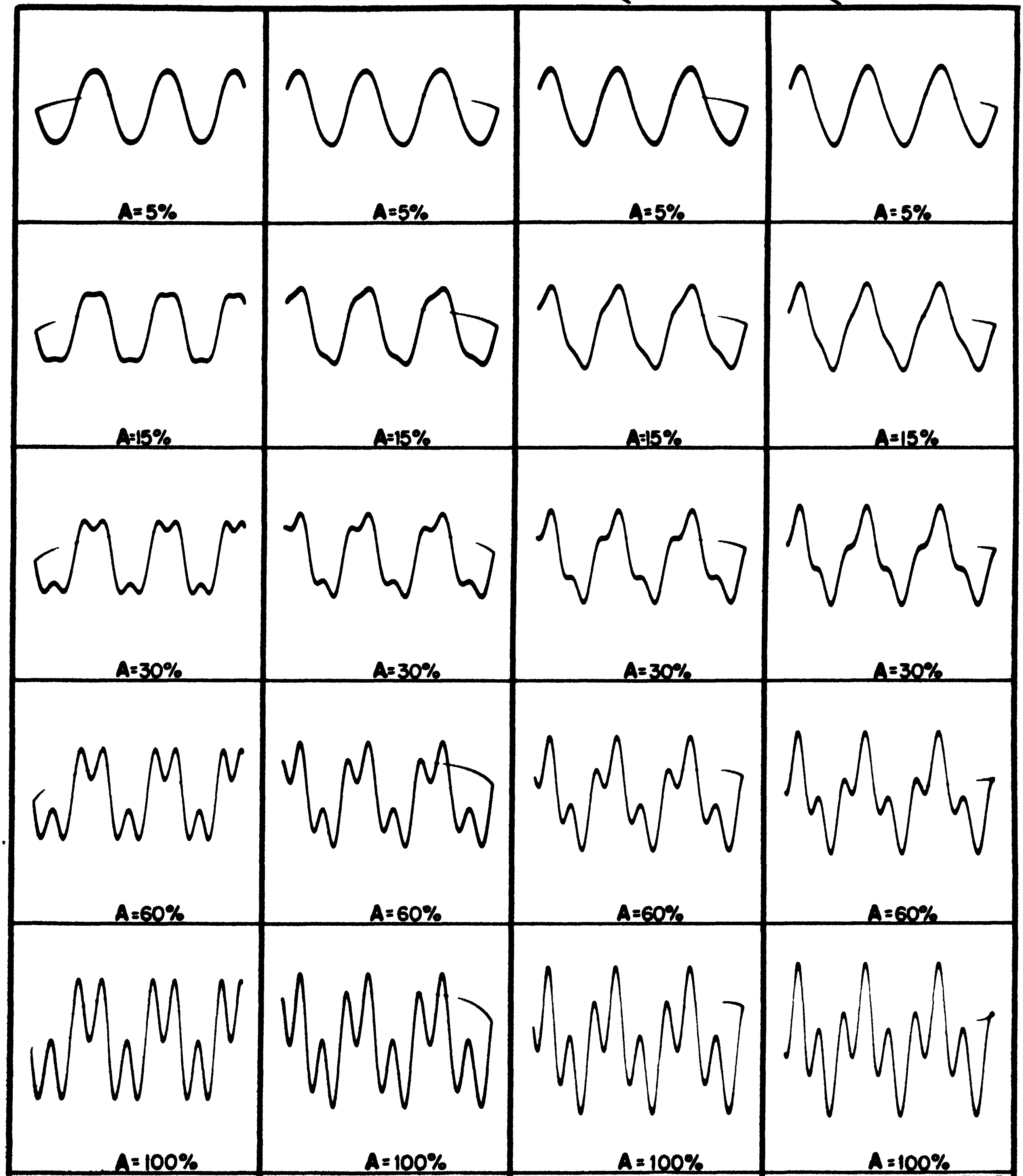

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3rd	0°	see A	3rd	+45°	see A	3rd	+90°	see A	3rd	+135°	see A

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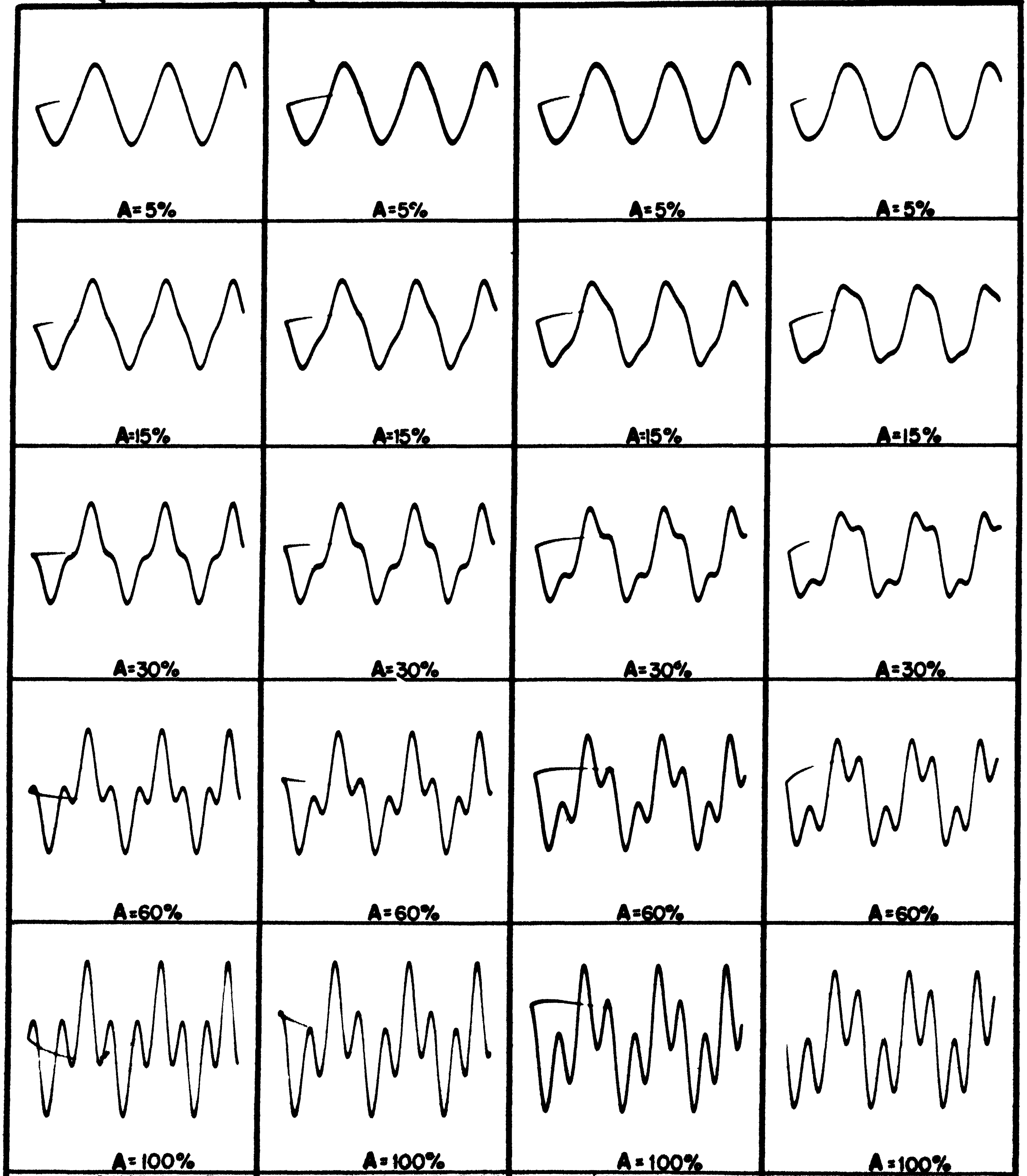

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Fund.	0°	100%
3rd	-90°	see A

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Fund.	0°	100%
3rd	-45°	see A

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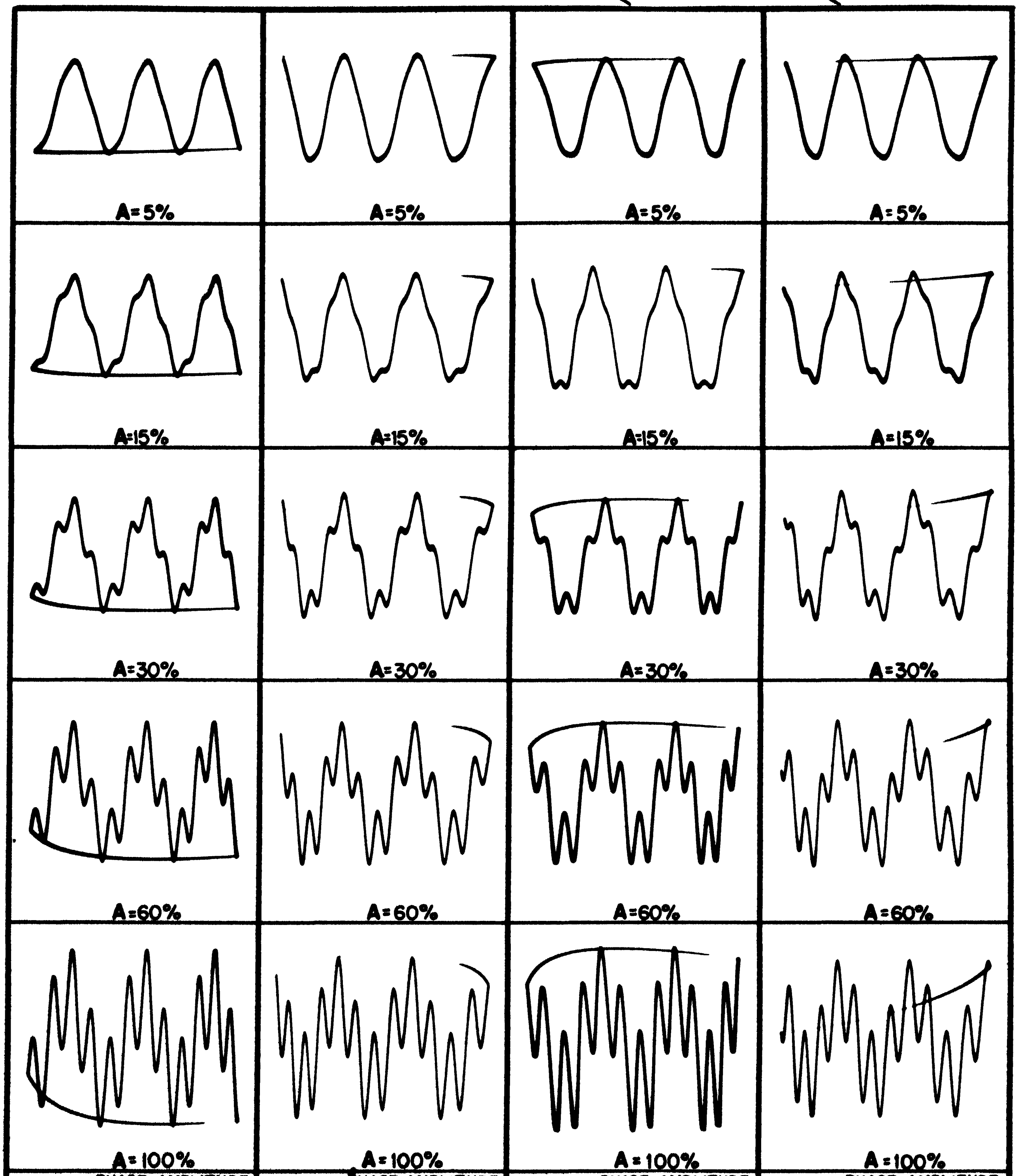

	PHASE	AMPLITUDE
Fund.	0°	100%
4th	0°	*see* A

	PHASE	AMPLITUDE
Fund.	0°	100%
4th	+45°	*see* A

	PHASE	AMPLITUDE
Fund.	0°	100%
4th	+90°	*see* A

	PHASE	AMPLITUDE
Fund.	0°	100%
4th	+135°	*see* A

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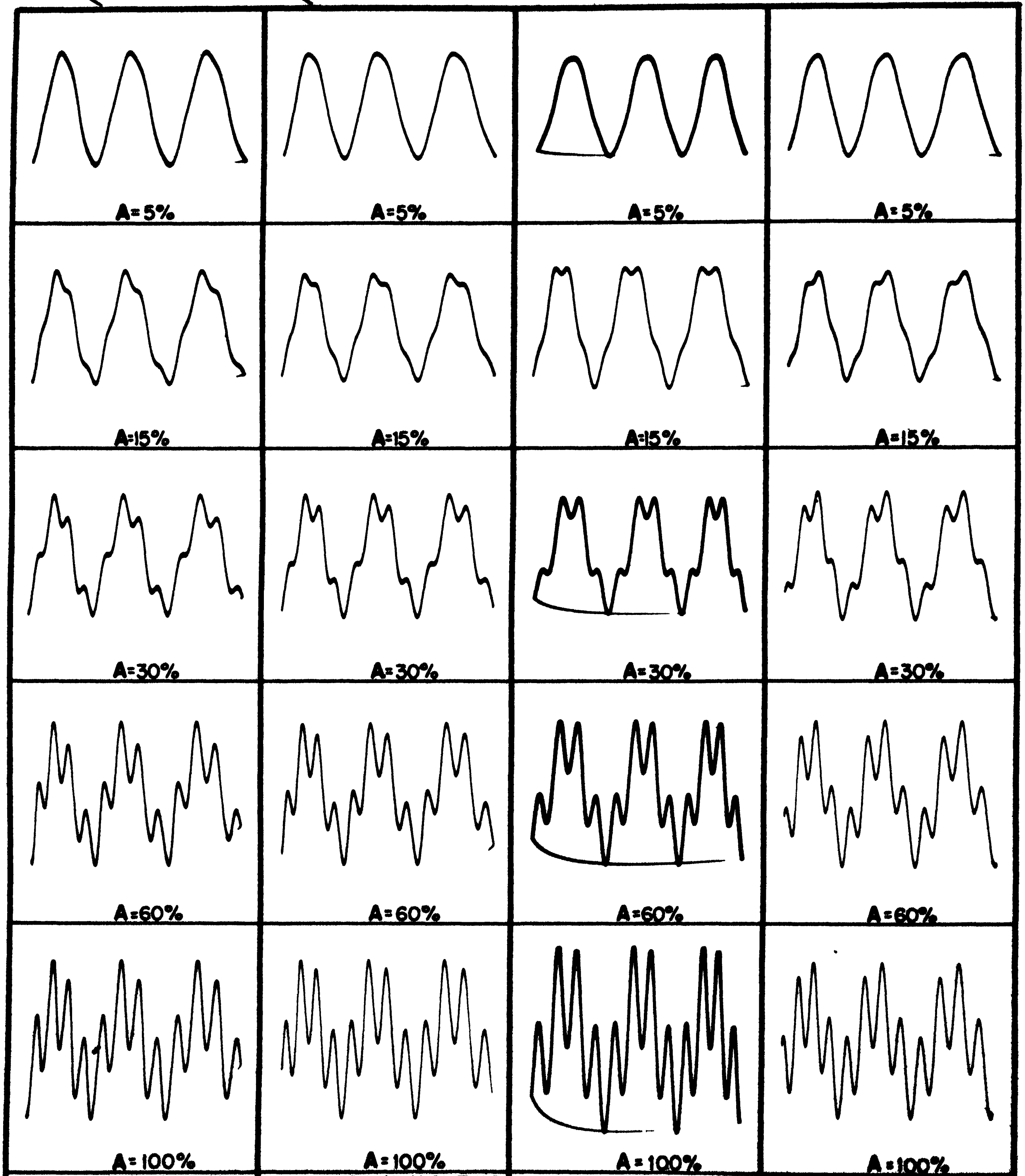

	PHASE	AMPLITUDE
Fund.	0°	100%
4th	180°	see A

	PHASE	AMPLITUDE
Fund.	0°	100%
4th	-135°	see A

	PHASE	AMPLITUDE
Fund.	0°	100%
4th	-90°	see A

	PHASE	AMPLITUDE
Fund.	0°	100%
4th	-45°	see A

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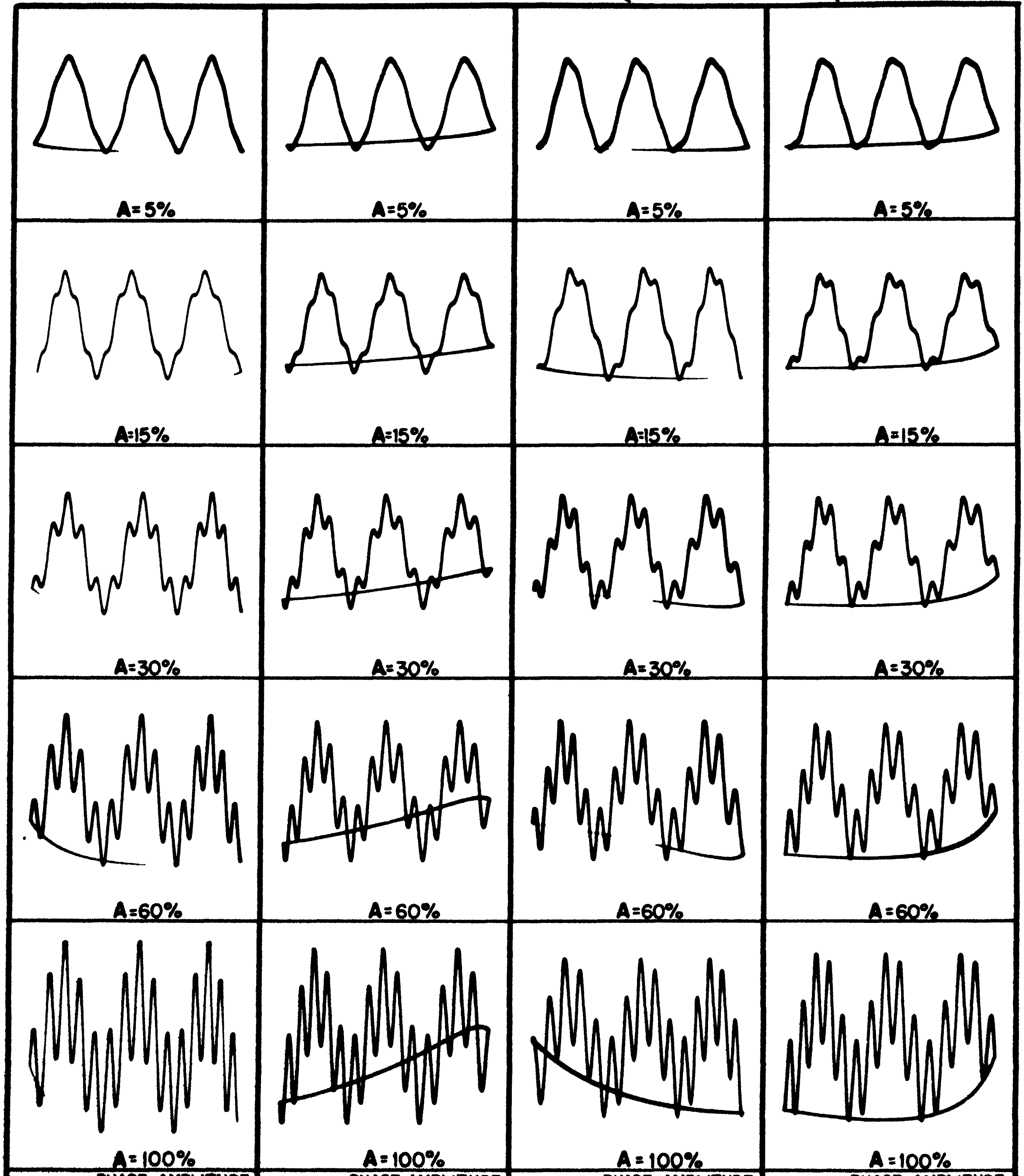


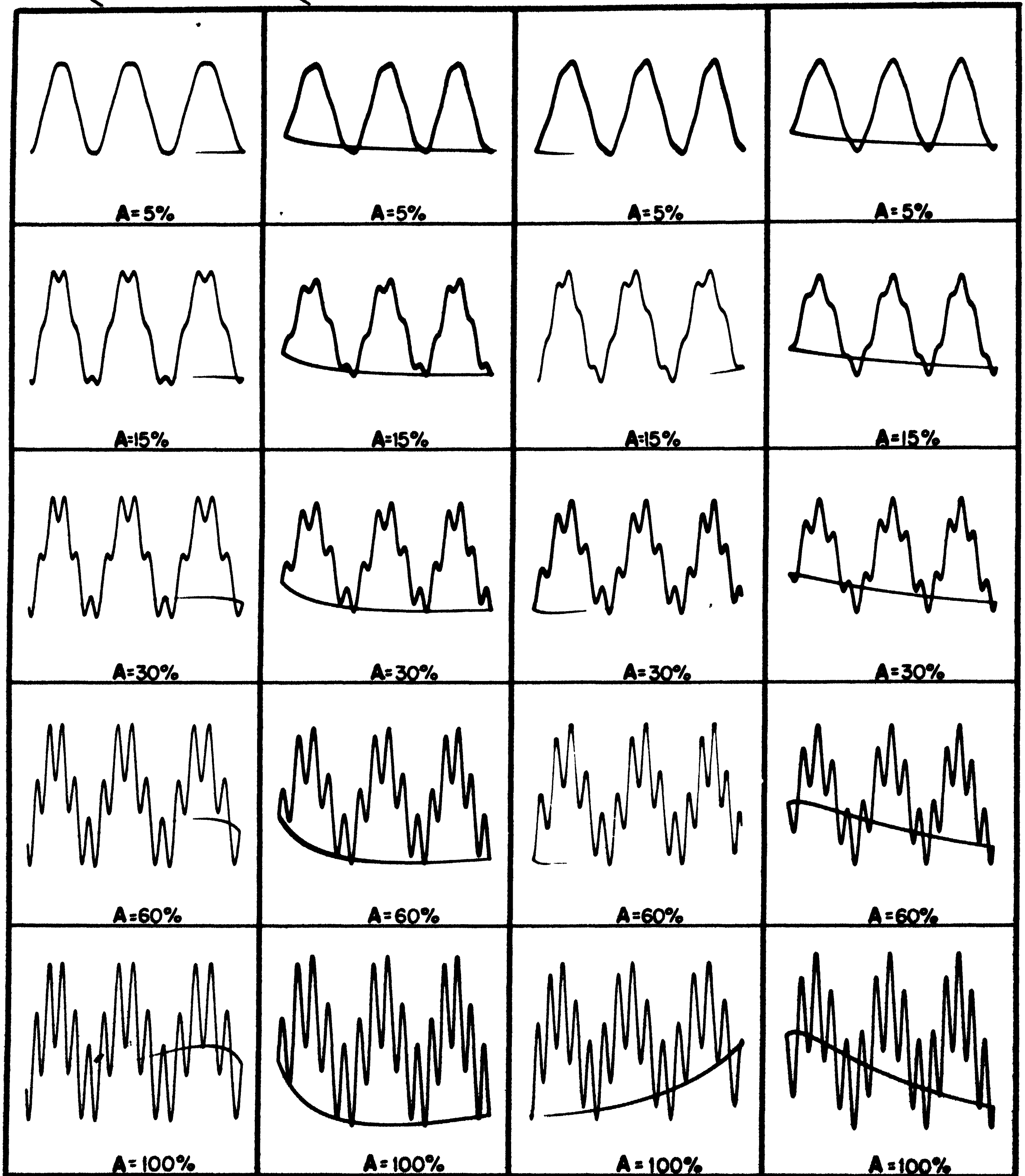
	PHASE	AMPLITUDE
Fund.	0°	100%
5th	0°	*see A*

	PHASE	AMPLITUDE
Fund.	0°	100%
5th	+45°	*see A*

	PHASE	AMPLITUDE
Fund.	0°	100%
5th	+90°	*see A*

	PHASE	AMPLITUDE
Fund.	0°	100%
5th	+135°	*see A*

$$\text{Fund.}\left\{\begin{matrix}0^\circ\\100\%\end{matrix}\right. + \text{5th}\left\{\begin{matrix}180^\circ \text{ to } -45^\circ\\ 5\% \text{ to } 100\%\end{matrix}\right.$$



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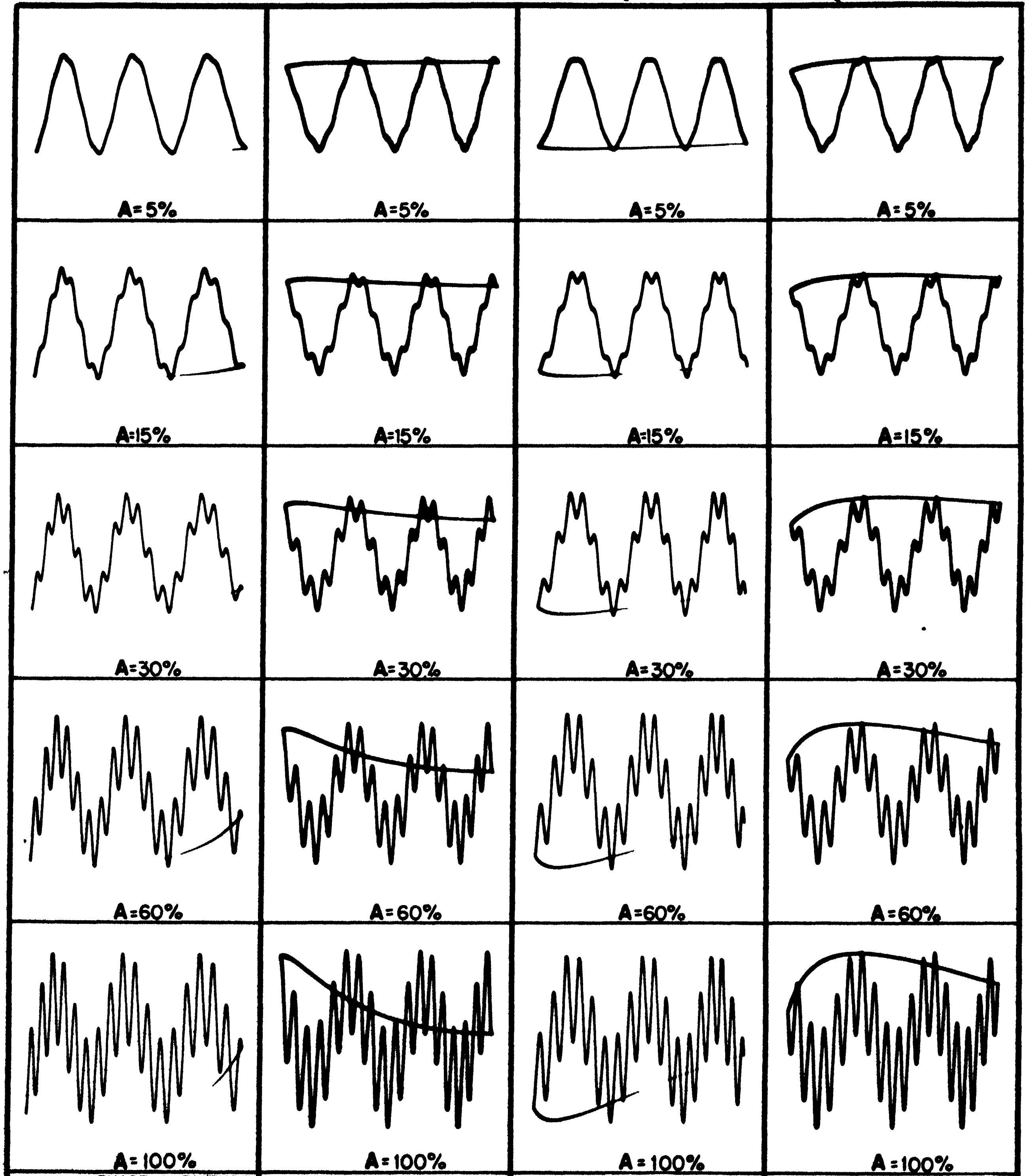
Fund.	0°	100%
5th	180°	*see* A

	PHASE	AMPLITUDE
Fund.	0°	100%
5th	-135°	*see* A

	PHASE	AMPLITUDE
Fund.	0°	100%
5th	-90°	*see* A

	PHASE	AMPLITUDE
Fund.	0°	100%
5th	-45°	*see* A

Fund. $\left\{ \begin{matrix} 0^\circ \\ 100\% \end{matrix} \right.$ **+ 6th** $\left\{ \begin{matrix} 0^\circ & \text{to} & +135^\circ \\ 5\% & \text{to} & 100\% \end{matrix} \right.$



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Fund.	0°	100%
6th	0°	*see A*

	PHASE	AMPLITUDE
Fund.	0°	100%
6th	+45°	*see A*

	PHASE	AMPLITUDE
Fund.	0°	100%
6th	+90°	*see A*

	PHASE	AMPLITUDE
Fund.	0°	100%
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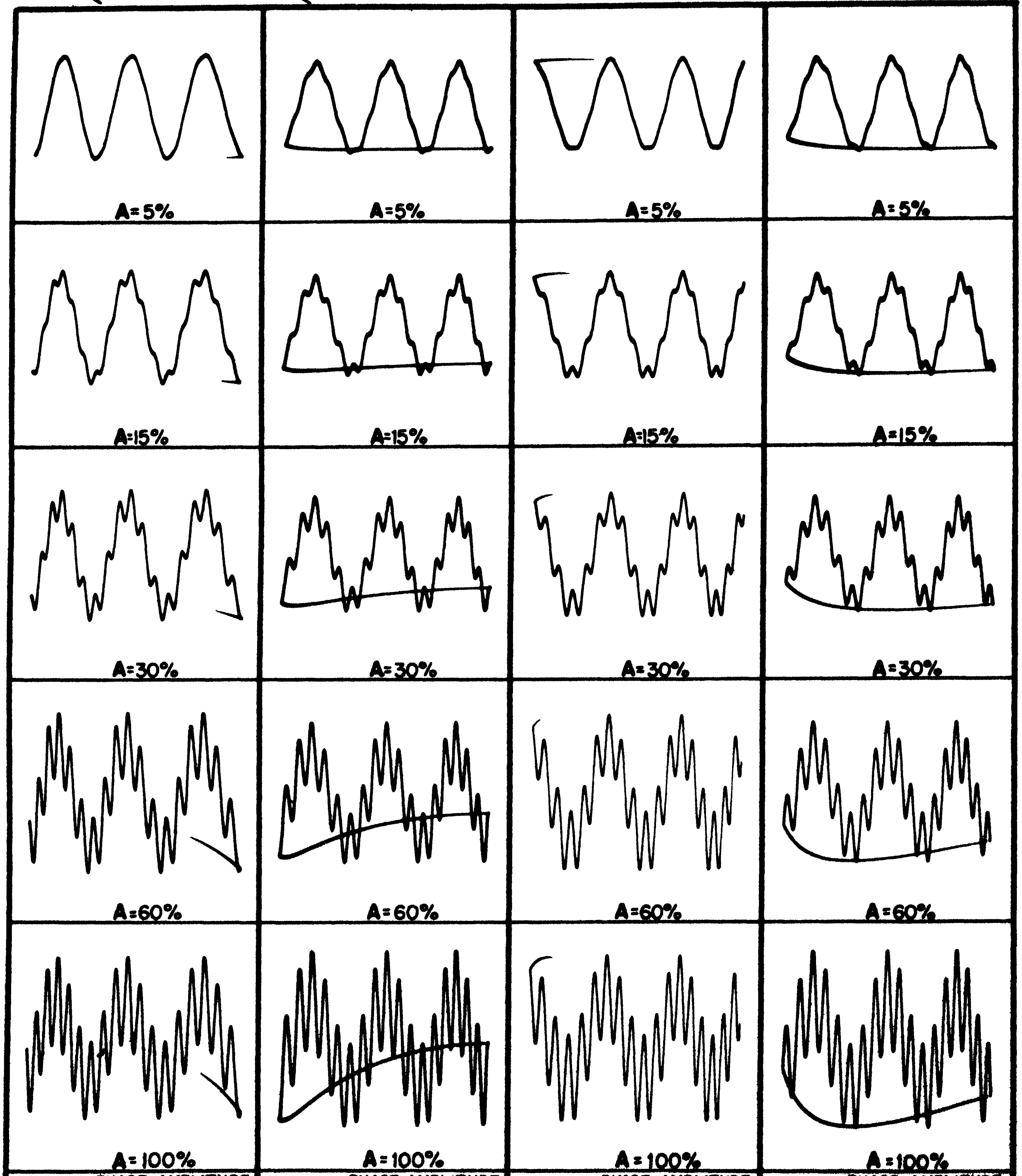

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	PHASE	AMPLITUDE
Fund.	0°	100%
6th	180°	see A

	PHASE	AMPLITUDE
Fund.	0°	100%
6th	−135°	see A

	PHASE	AMPLITUDE
Fund.	0°	100%
6th	−90°	see A

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Fund.	0°	100%
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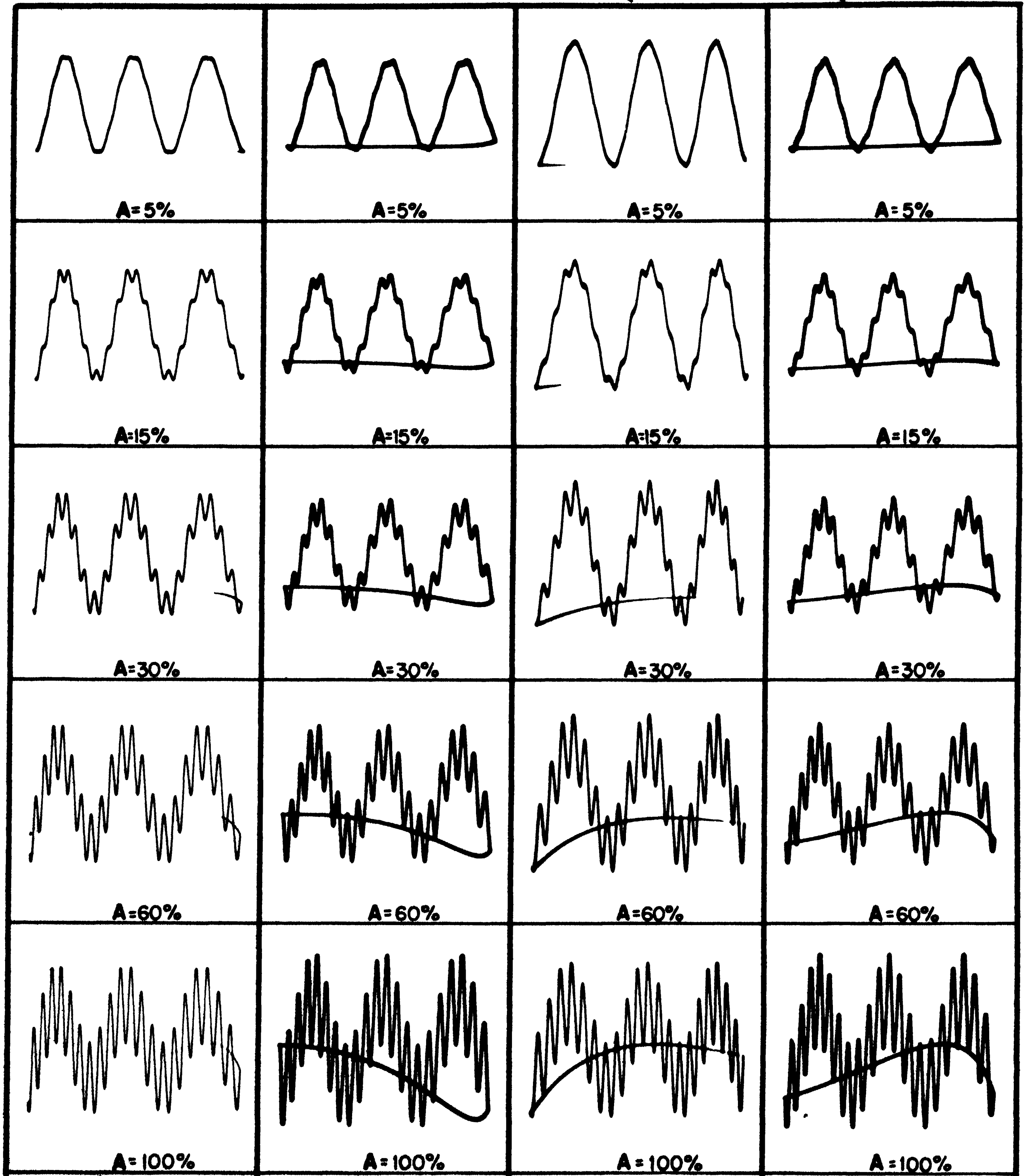


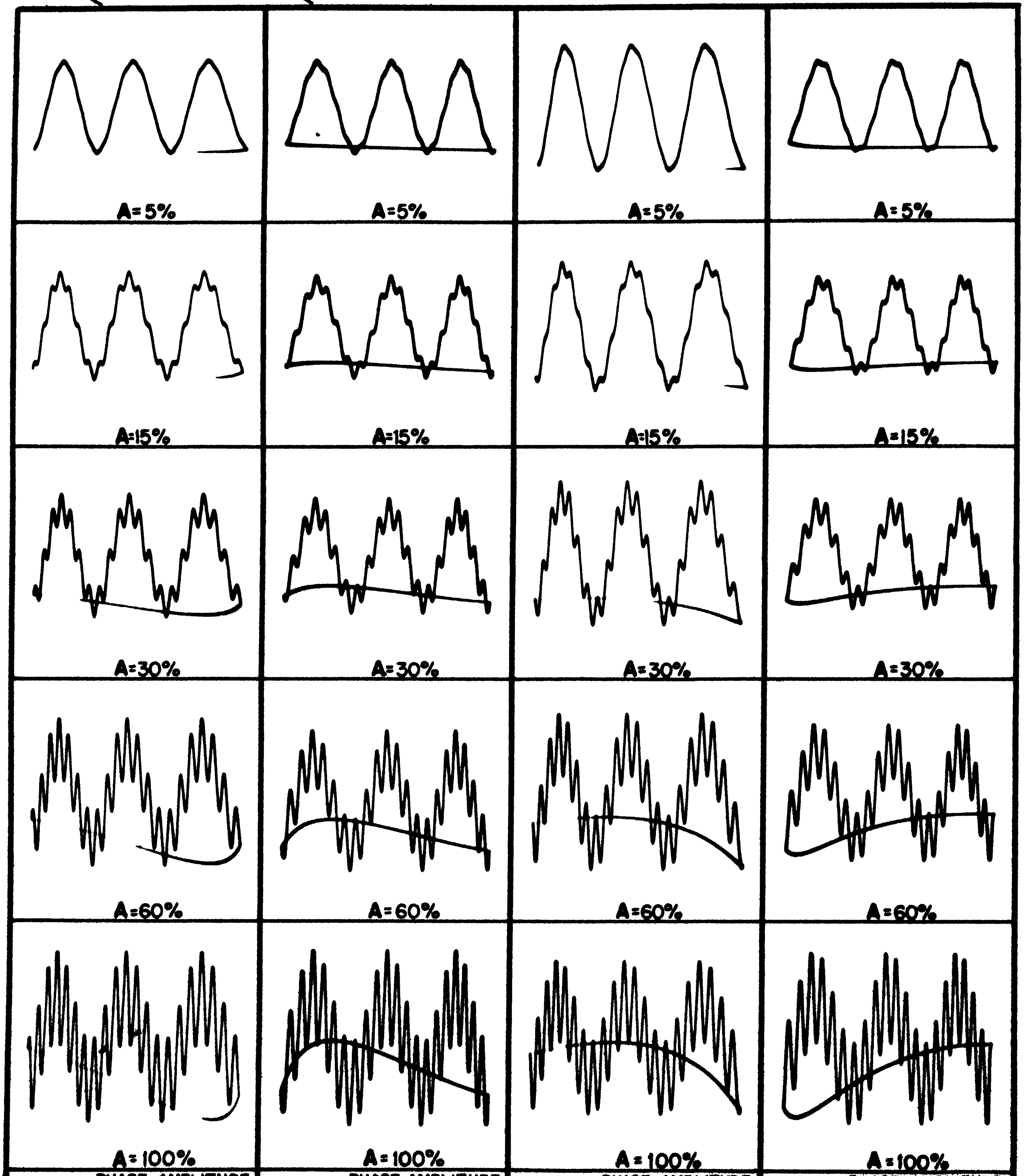
	PHASE	AMPLITUDE
Fund.	0°	100°
7th	0°	see A

	PHASE	AMPLITUDE
Fund.	0°	100%
7th	+45°	see A

	PHASE	AMPLITUDE
Fund.	0°	100%
7th	+90°	see A

	PHASE	AMPLITUDE
Fund.	0°	100%
7th	+135°	see A

Fund. { 0° 100 } + 7th { 180° to −45° 5% to 100% }



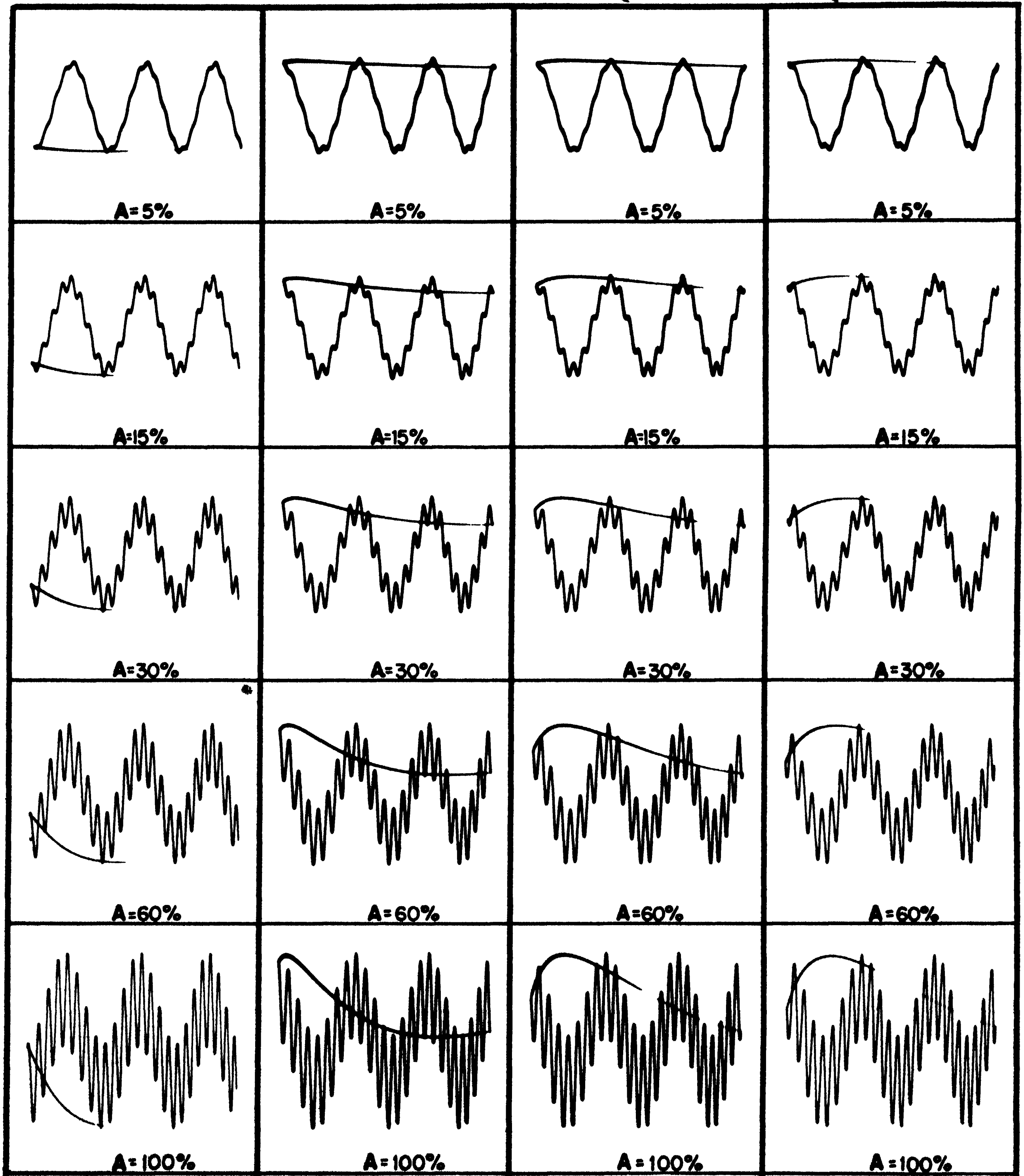
	PHASE	AMPLITUDE
Fund.	0°	100%
7th	180°	see A

	PHASE	AMPLITUDE
Fund.	0°	100%
7th	−135°	see A

	PHASE	AMPLITUDE
Fund.	0°	100%
7th	−90°	see A

	PHASE	AMPLITUDE
Fund.	0°	100%
7th	−45°	see A

Fund. $\left\{\begin{matrix} 0° \\ 100\% \end{matrix}\right.$ + 8th $\left\{\begin{matrix} 0° & \text{to} & +135° \\ 5\% & \text{to} & 100\% \end{matrix}\right.$



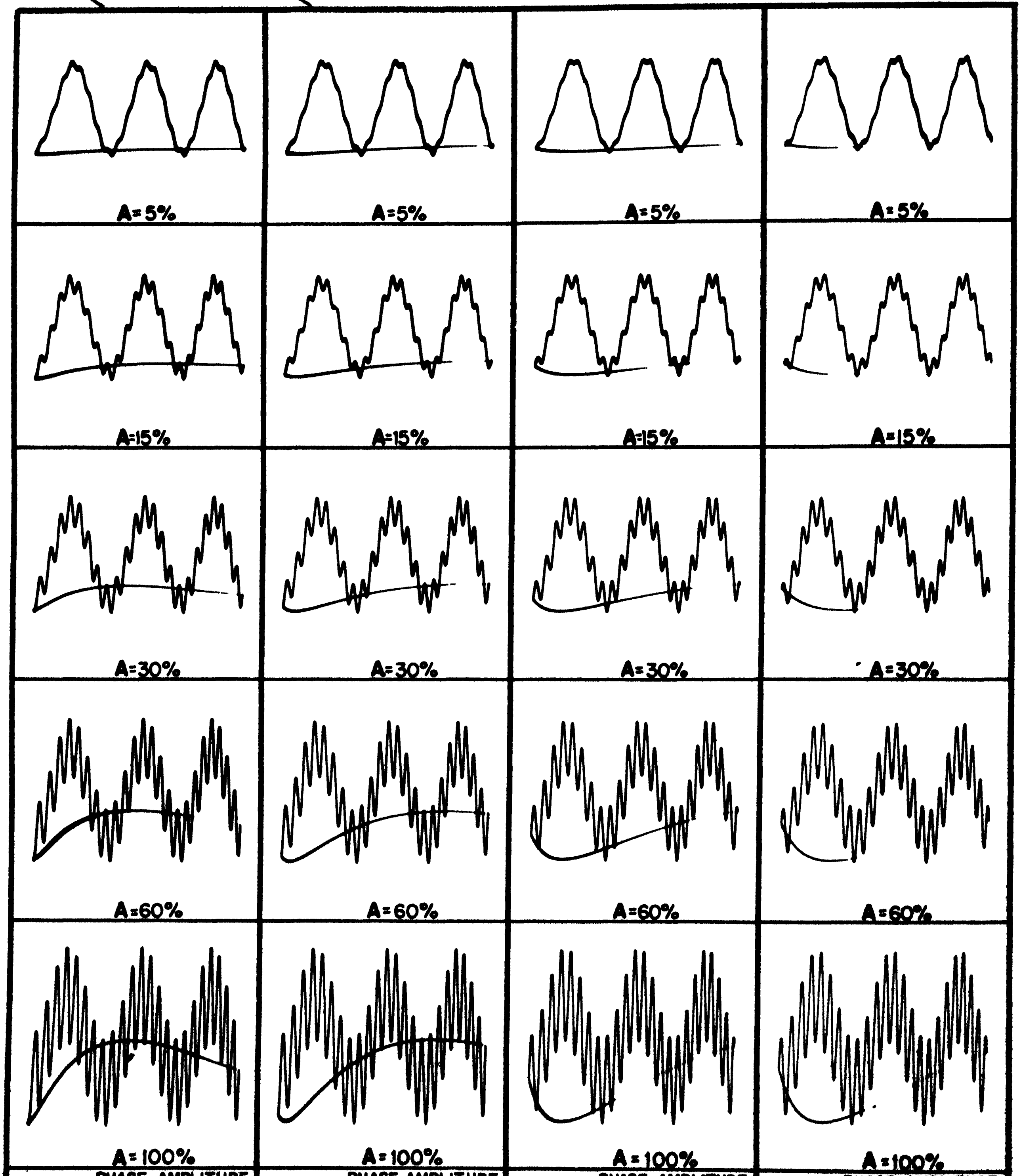
	PHASE	AMPLITUDE
Fund.	0°	100%
8th	0°	see A

	PHASE	AMPLITUDE
Fund.	0°	100%
8th	+45°	see A

	PHASE	AMPLITUDE
Fund.	0°	100%
8th	+90°	see A

	PHASE	AMPLITUDE
Fund.	0°	100%
8th	+135°	see A

$$\text{Fund.}\begin{cases}0^\circ\\100\%\end{cases} + \text{8th}\begin{cases}180^\circ \text{ to } -45^\circ\\5\% \text{ to } 100\%\end{cases}$$



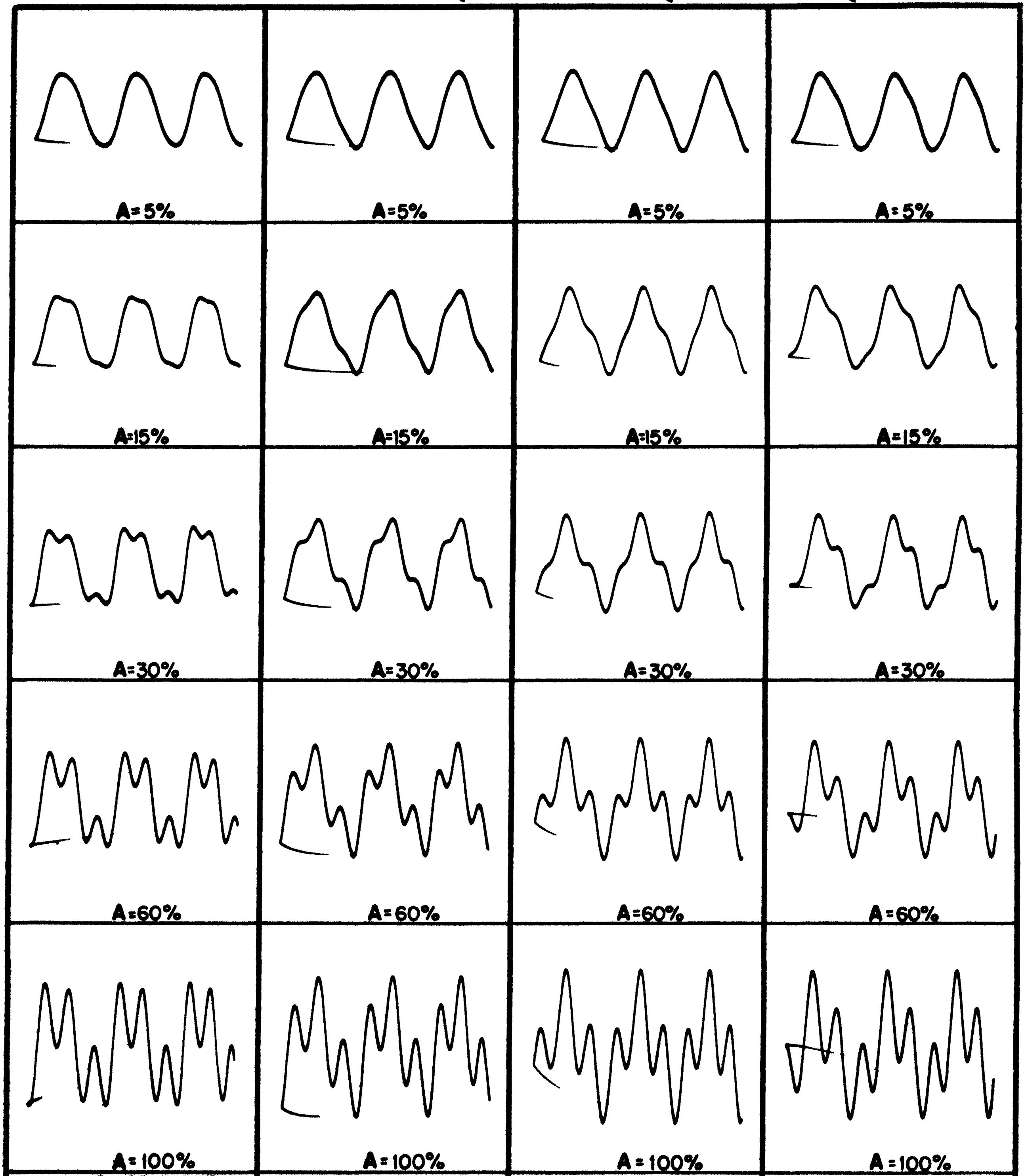
	PHASE	AMPLITUDE
Fund.	0°	100%
8th	180°	see A

	PHASE	AMPLITUDE
Fund.	0°	100%
8th	-135°	see A

	PHASE	AMPLITUDE
Fund.	0°	100%
8th	-90°	see A

	PHASE	AMPLITUDE
Fund.	0°	100%
8th	-45°	see A

Fund. {0° 100% + 2nd {0° 5% + 3rd {0° to −90° 5% to 100%



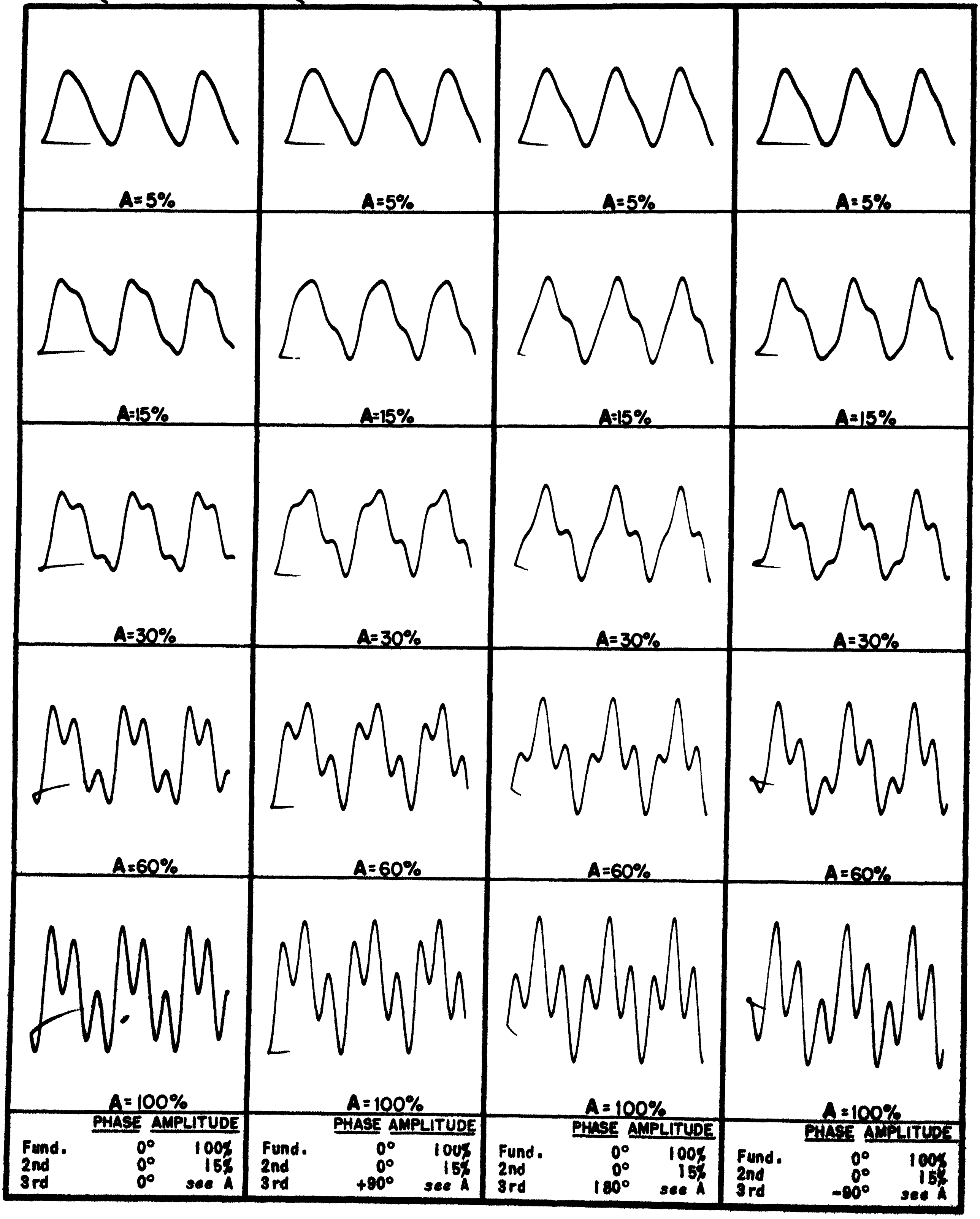
	PHASE	AMPLITUDE
Fund.	0°	100%
2nd	0°	5%
3rd	0°	see A

	PHASE	AMPLITUDE
Fund.	0°	100%
2nd	0°	5%
3rd	+90°	see A

	PHASE	AMPLITUDE
Fund.	0°	100%
2nd	0°	5%
3rd	180°	see A

	PHASE	AMPLITUDE
Fund.	0°	100%
2nd	0°	5%
3rd	-90°	see A

Fund. $\left\{\begin{matrix}0^\circ \\ 100\%\end{matrix}\right.$ + 2nd $\left\{\begin{matrix}0^\circ \\ 15\%\end{matrix}\right.$ + 3rd $\left\{\begin{matrix}0^\circ \text{ to } -90^\circ \\ 5\% \text{ to } 100\%\end{matrix}\right.$



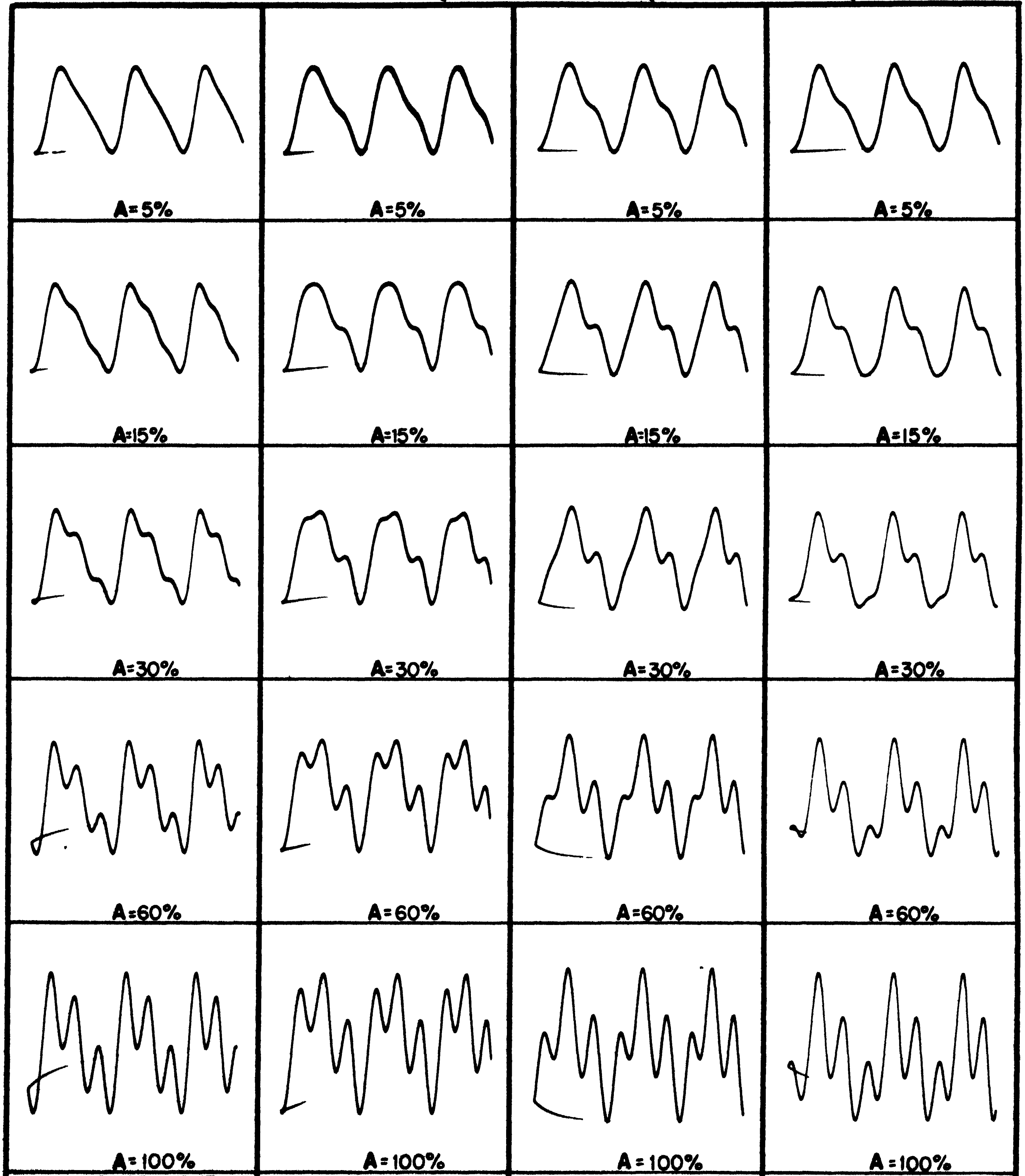
	PHASE	AMPLITUDE
Fund.	0°	100%
2nd	0°	15%
3rd	0°	*see* A

	PHASE	AMPLITUDE
Fund.	0°	100%
2nd	0°	15%
3rd	+90°	*see* A

	PHASE	AMPLITUDE
Fund.	0°	100%
2nd	0°	15%
3rd	180°	*see* A

	PHASE	AMPLITUDE
Fund.	0°	100%
2nd	0°	15%
3rd	-90°	*see* A

Fund. $\left\{\begin{matrix} 0° \\ 100\% \end{matrix}\right.$ + 2nd $\left\{\begin{matrix} 0° \\ 30\% \end{matrix}\right.$ + 3rd $\left\{\begin{matrix} 0° \text{ to } -90° \\ 5\% \text{ to } 100\% \end{matrix}\right.$



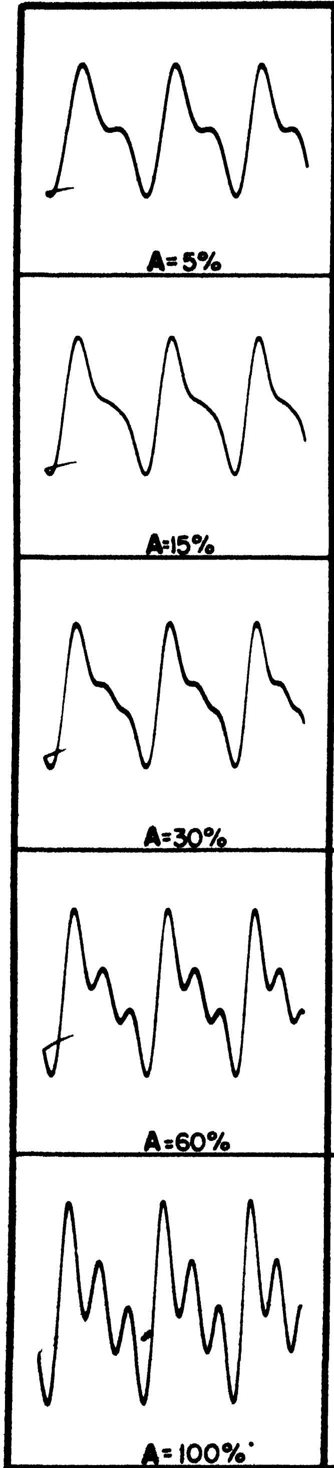
	PHASE	AMPLITUDE
Fund.	0°	100%
2nd	0°	30%
3rd	0°	*see* A

	PHASE	AMPLITUDE
Fund.	0°	100%
2nd	0°	30%
3rd	+90°	*see* A

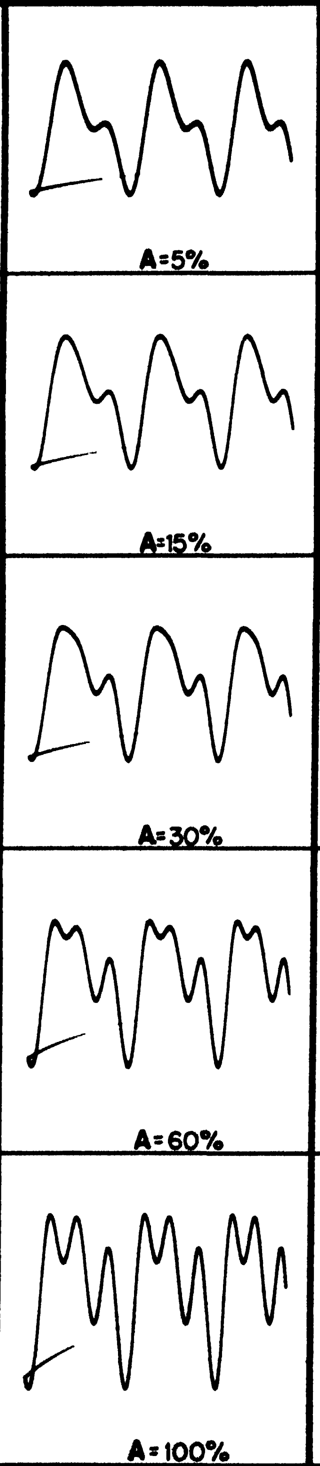
	PHASE	AMPLITUDE
Fund.	0°	100%
2nd	0°	30%
3rd	180°	*see* A

	PHASE	AMPLITUDE
Fund.	0°	100%
2nd	0°	30%
3rd	-90°	*see* A

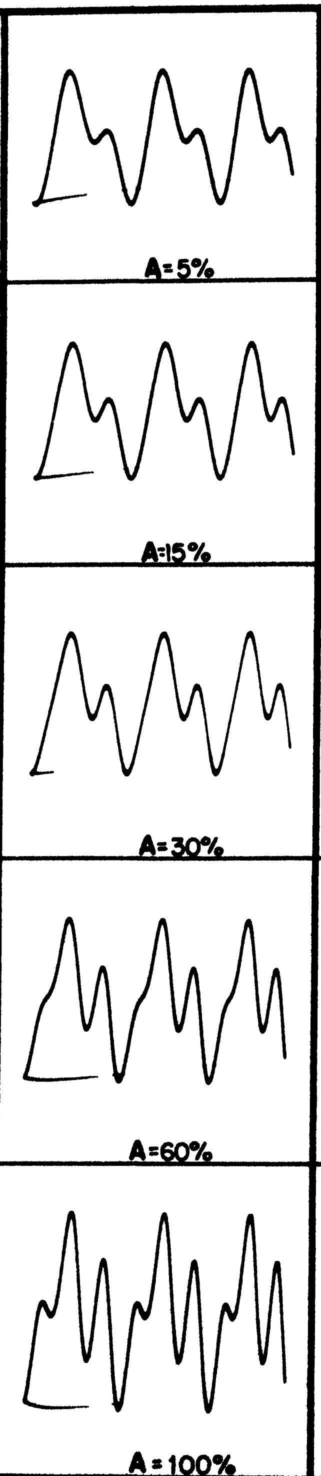
$$\text{Fund.}\left\{\begin{matrix}0^\circ\\ 100\%\end{matrix}\right. + \text{2nd}\left\{\begin{matrix}0^\circ\\ 60\%\end{matrix}\right. + \text{3rd}\left\{\begin{matrix}0^\circ \text{ to } -90^\circ\\ 5\% \text{ to } 100\%\end{matrix}\right.$$



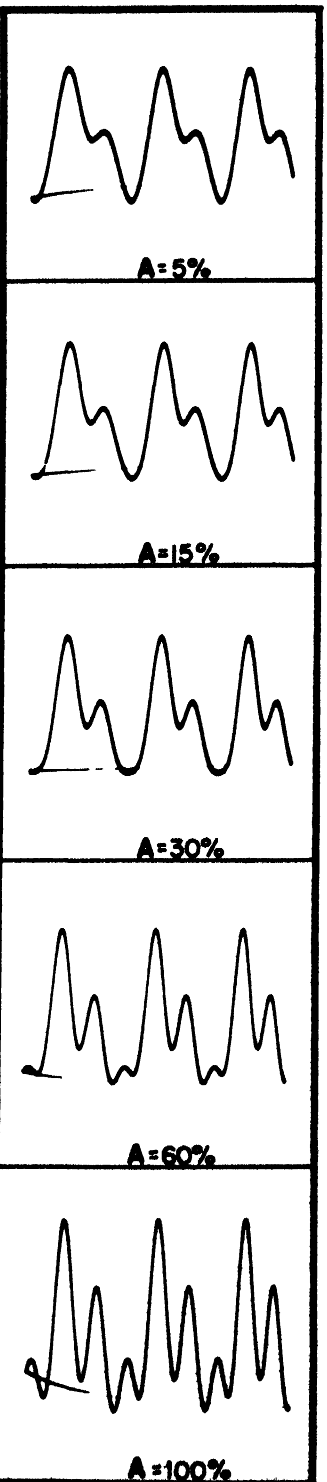
	PHASE	AMPLITUDE
Fund.	0°	100%
2nd	0°	60%
3rd	0°	see A



	PHASE	AMPLITUDE
Fund.	0°	100%
2nd	0°	60%
3rd	+90°	see A

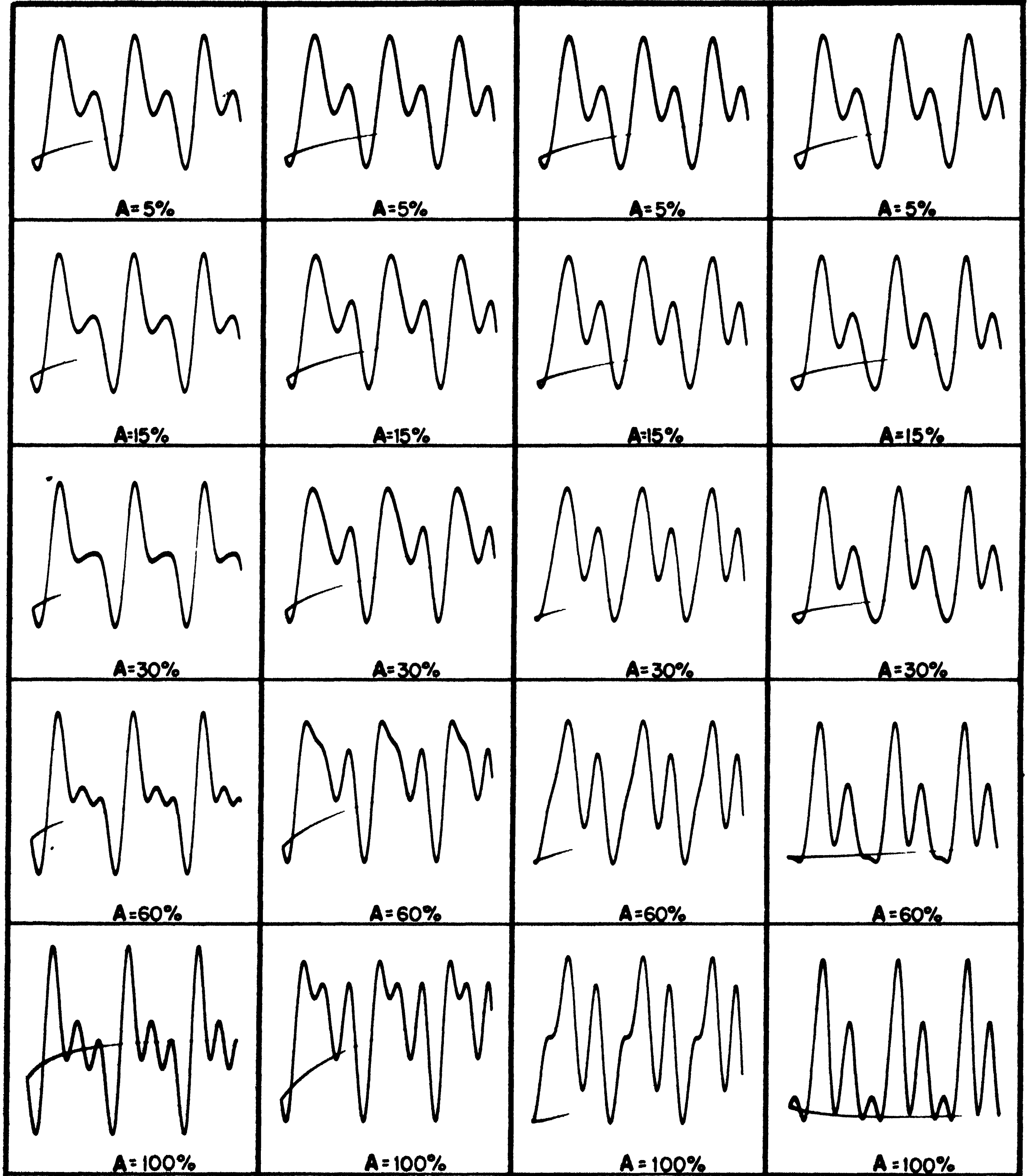


	PHASE	AMPLITUDE
Fund.	0°	100%
2nd	0°	60%
3rd	180°	see A



	PHASE	AMPLITUDE
Fund.	0°	100%
2nd	0°	60%
3rd	-90°	see A

$$\text{Fund.}\left\{\begin{matrix}0^\circ\\100\%\end{matrix}\right. + \ 2\text{nd}\left\{\begin{matrix}0^\circ\\100\%\end{matrix}\right. + \ 3\text{rd}\left\{\begin{matrix}0^\circ \text{ to } -90^\circ\\5\% \text{ to } 100\%\end{matrix}\right.$$



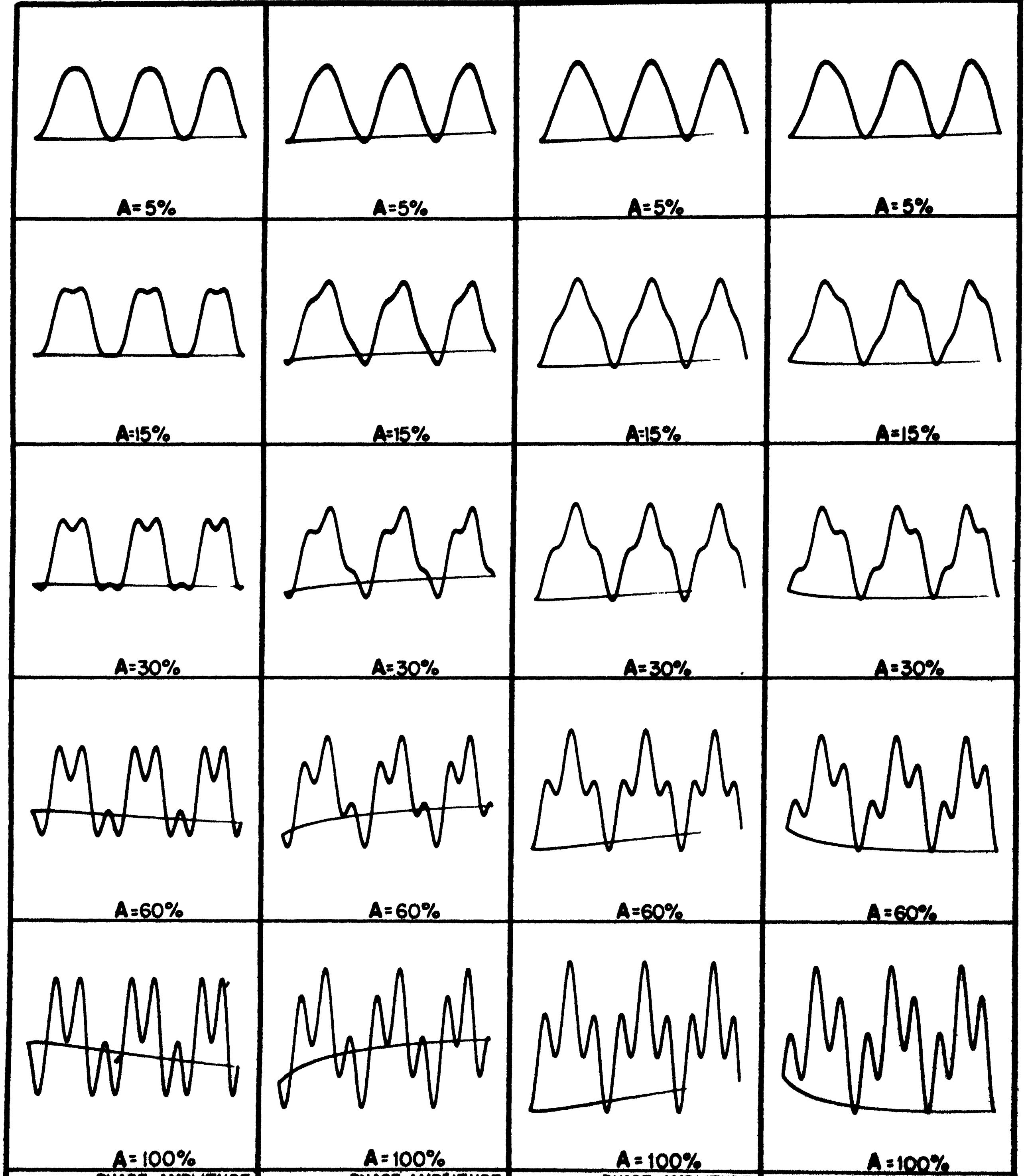
	PHASE	AMPLITUDE
Fund.	0°	100%
2nd	0°	100%
3rd	0°	see A

	PHASE	AMPLITUDE
Fund.	0°	100%
2nd	0°	100%
3rd	+90°	see A

	PHASE	AMPLITUDE
Fund.	0°	100%
2nd	0°	100%
3rd	180°	see A

	PHASE	AMPLITUDE
Fund.	0°	100%
2nd	0°	100%
3rd	-90°	see A

$$\text{Fund.}\begin{cases}0^\circ\\100\%\end{cases} + \text{2nd}\begin{cases}+90^\circ\\5\%\end{cases} + \text{3rd}\begin{cases}0^\circ \text{ to } -90^\circ\\5\% \text{ to } 100\%\end{cases}$$



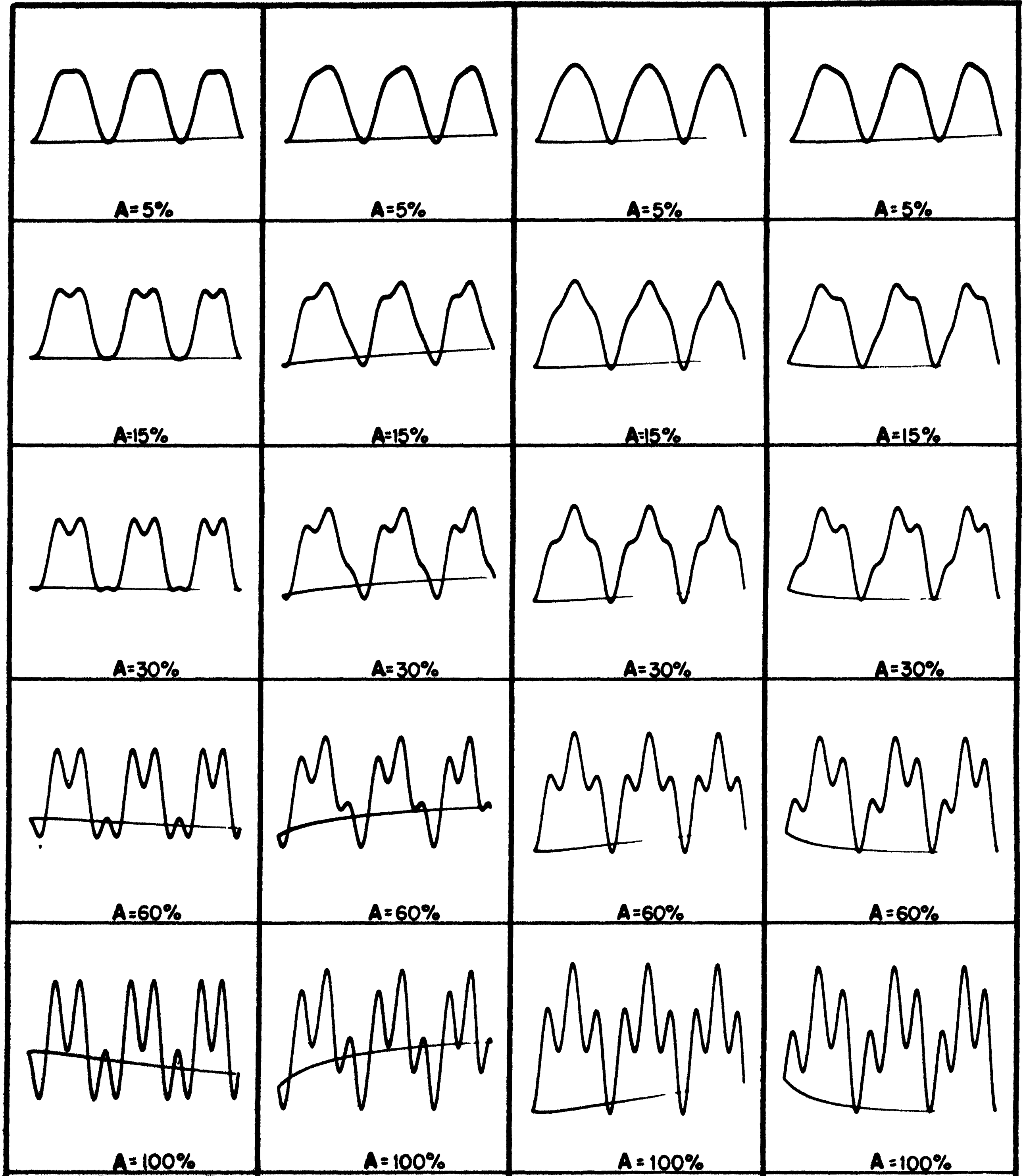
	PHASE	AMPLITUDE
Fund.	0°	100%
2nd	+90°	5%
3rd	0°	see A

	PHASE	AMPLITUDE
Fund.	0°	100%
2nd	+90°	5%
3rd	+90°	see A

	PHASE	AMPLITUDE
Fund.	0°	100%
2nd	+90°	5%
3rd	180°	see A

	PHASE	AMPLITUDE
Fund.	0°	100%
2nd	+90°	5%
3rd	-90°	see A

$$\text{Fund.}\left\{\begin{matrix}0^\circ\\100\%\end{matrix}\right. + \text{2nd}\left\{\begin{matrix}+90^\circ\\15\%\end{matrix}\right. + \text{3rd}\left\{\begin{matrix}0^\circ & \text{to} & -90^\circ\\5\% & \text{to} & 100\%\end{matrix}\right.$$



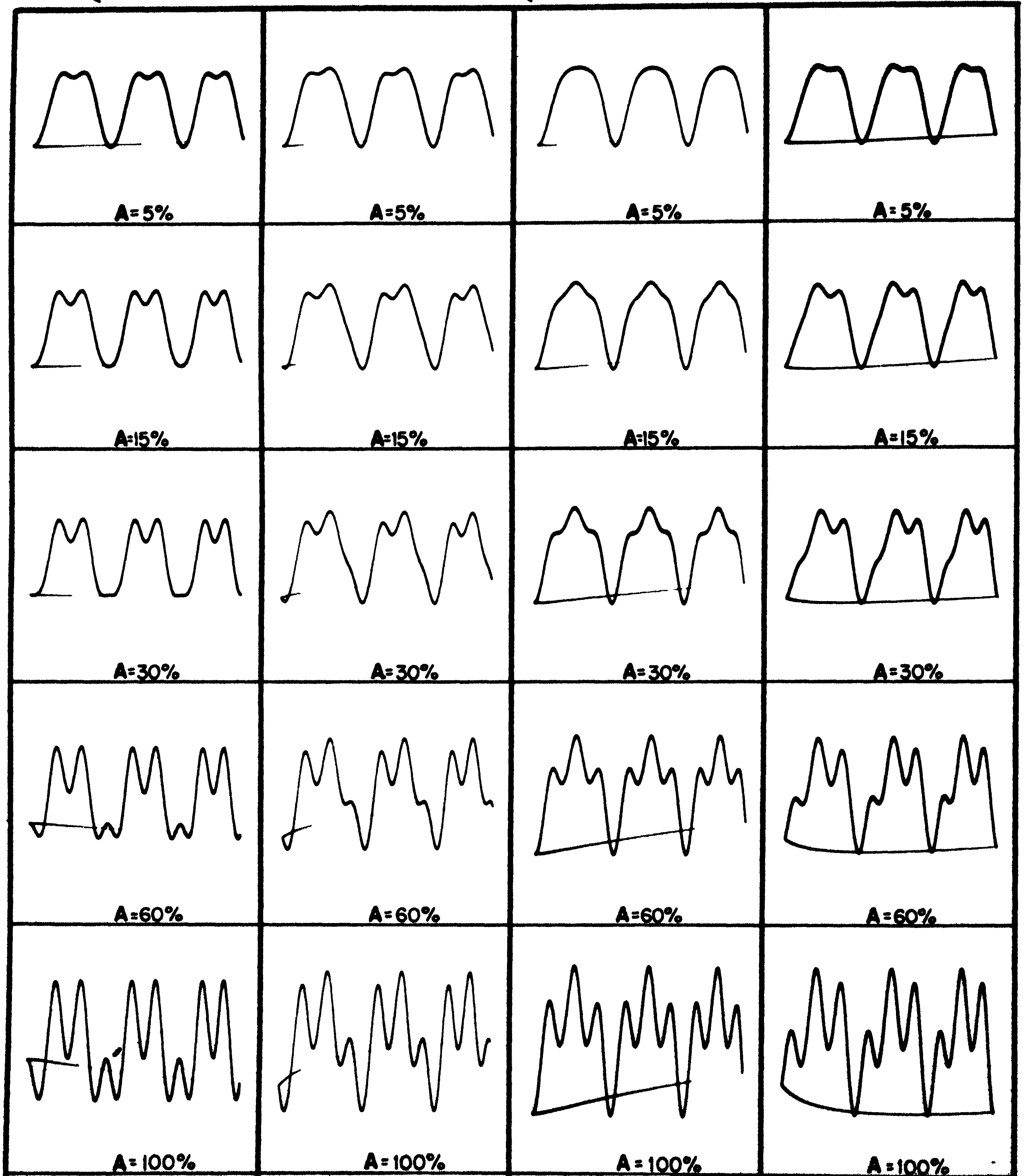
	PHASE	AMPLITUDE
Fund.	0°	100%
2nd	+90°	15%
3rd	0°	*see* A

	PHASE	AMPLITUDE
Fund.	0°	100%
2nd	+90°	15%
3rd	+90°	*see* A

	PHASE	AMPLITUDE
Fund.	0°	100%
2nd	+90°	15%
3rd	180°	*see* A

	PHASE	AMPLITUDE
Fund.	0°	100%
2nd	+90°	15%
3rd	-90°	*see* A

Fund. { 0° 100% + 2nd { +90° 30% + 3rd { 0° to -90° 5% to 100%



A=5% A=5% A=5% A=5%

A=15% A=15% A=15% A=15%

A=30% A=30% A=30% A=30%

A=60% A=60% A=60% A=60%

A=100% A=100% A=100% A=100%

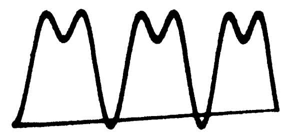
	PHASE	AMPLITUDE
Fund.	0°	100%
2nd	+90°	30%
3rd	0°	see A

	PHASE	AMPLITUDE
Fund.	0°	100%
2nd	+90°	30%
3rd	+90°	see A

	PHASE	AMPLITUDE
Fund.	0°	100%
2nd	+90°	30%
3rd	180°	see A

	PHASE	AMPLITUDE
Fund.	0°	100%
2nd	+90°	30%
3rd	-90°	see A

$$\text{Fund.}\begin{cases}0^\circ\\100\%\end{cases} + \text{2nd}\begin{cases}+90^\circ\\60\%\end{cases} + \text{3rd}\begin{cases}0^\circ \text{ to } -90^\circ\\5\% \text{ to } 100\%\end{cases}$$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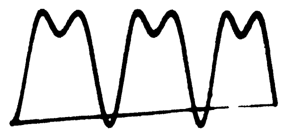
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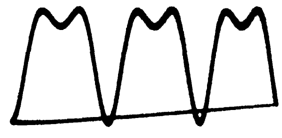
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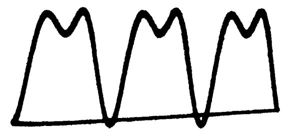
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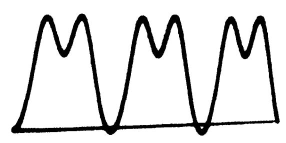
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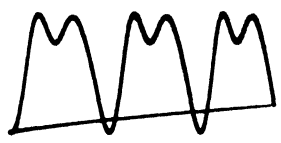
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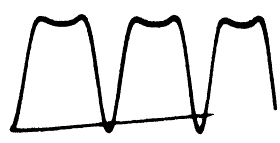
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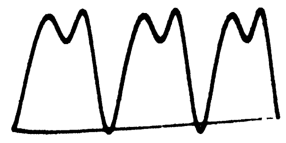
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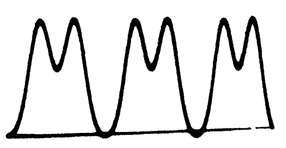
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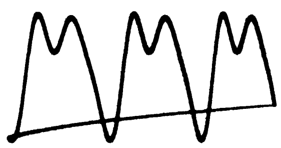
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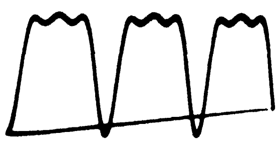
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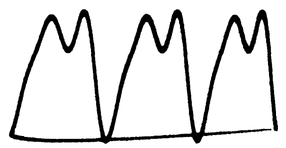
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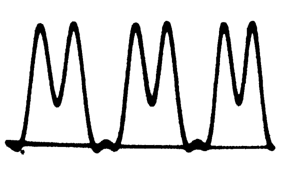
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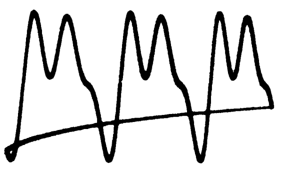
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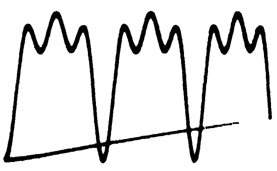
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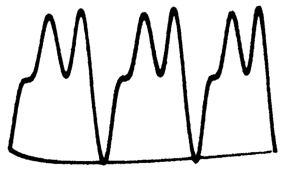
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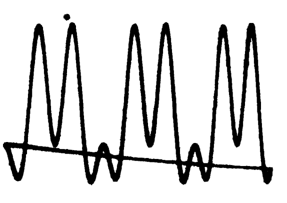
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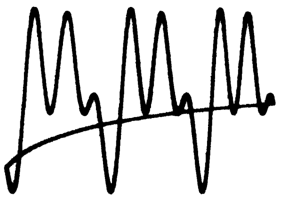
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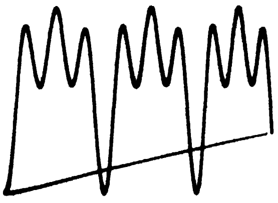
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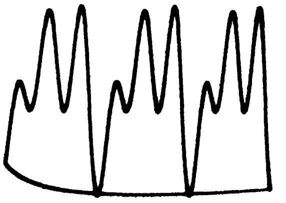

A=100%

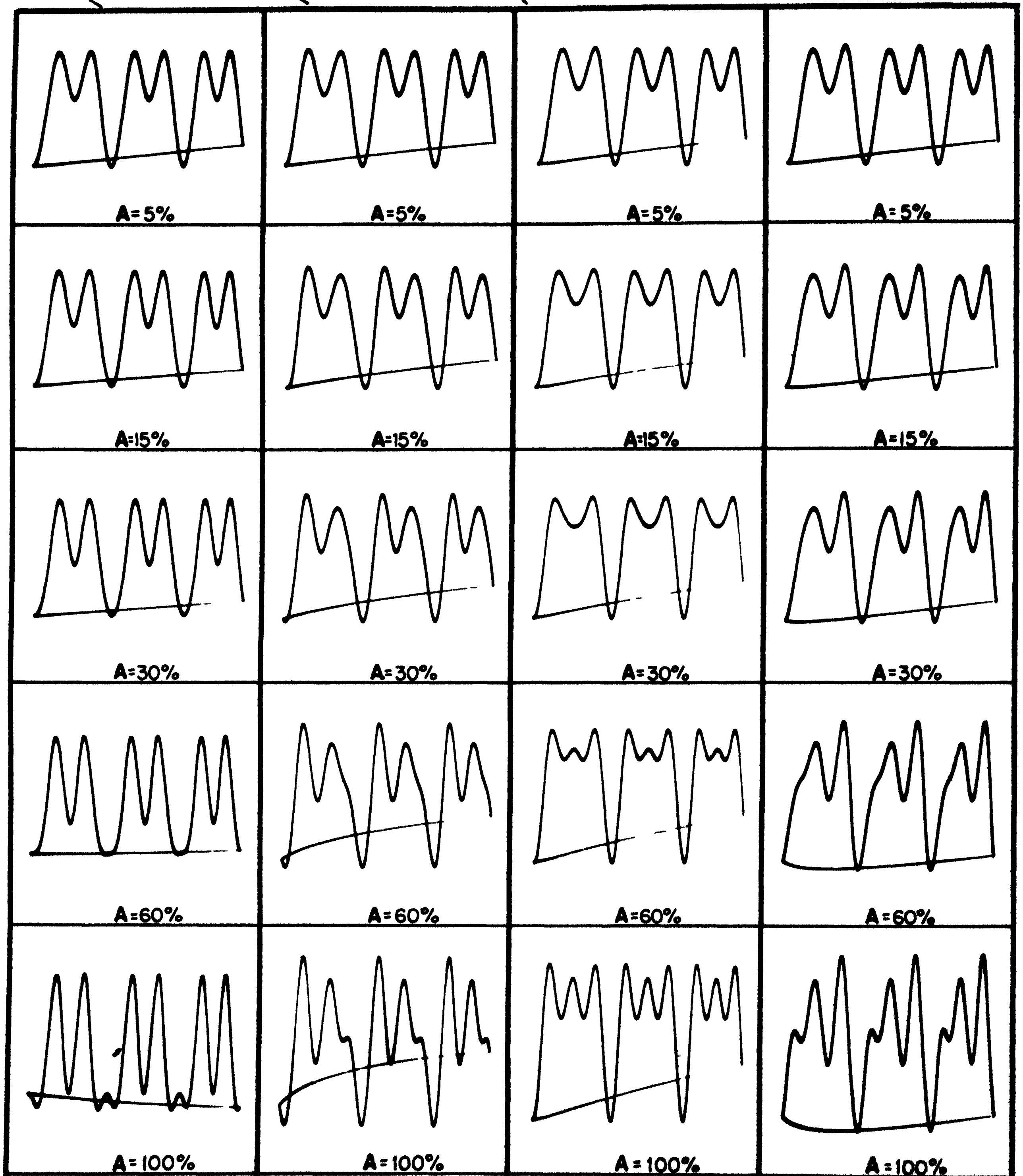
	PHASE	AMPLITUDE
Fund.	0°	100%
2nd	+90°	60%
3rd	0°	see A

	PHASE	AMPLITUDE
Fund.	0°	100%
2nd	+90°	60%
3rd	+90°	see A

	PHASE	AMPLITUDE
Fund.	0°	100%
2nd	+90°	60%
3rd	180°	see A

	PHASE	AMPLITUDE
Fund.	0°	100%
2nd	+90°	60%
3rd	-90°	see A

Fund. {0° 100%} + 2nd {+90° 100%} + 3rd {0° to −90° 5% to 100%}



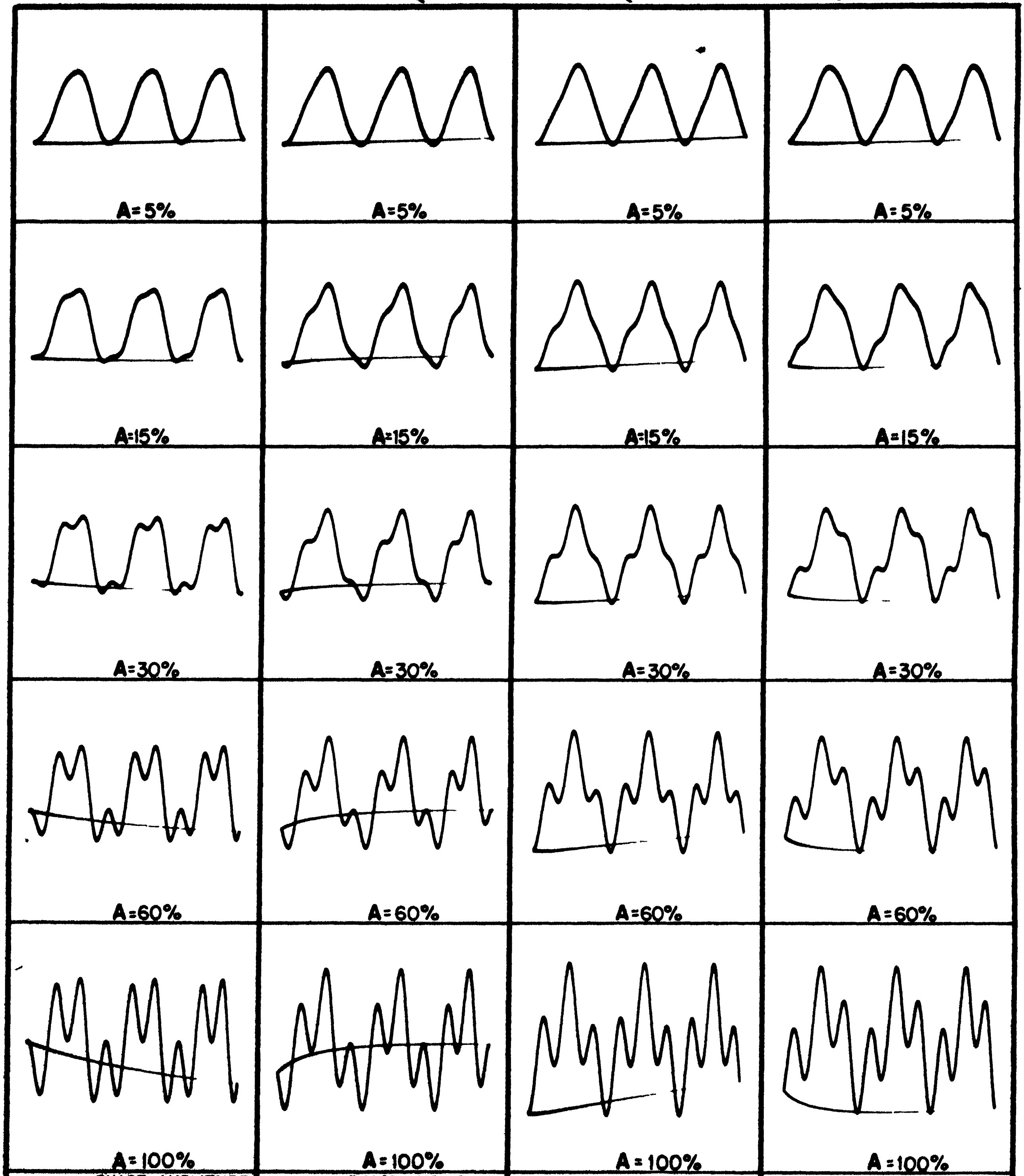
	PHASE	AMPLITUDE
Fund.	0°	100%
2nd	+90°	100%
3rd	0°	*see A*

	PHASE	AMPLITUDE
Fund.	0°	100%
2nd	+90°	100%
3rd	+90°	*see A*

	PHASE	AMPLITUDE
Fund.	0°	100%
2nd	+90°	100%
3rd	180°	*see A*

	PHASE	AMPLITUDE
Fund.	0°	100%
2nd	+90°	100%
3rd	-90°	*see A*

Fund. {0° 100%} + 2nd {180° 5%} + 3rd {0° to −90° 5% to 100%}



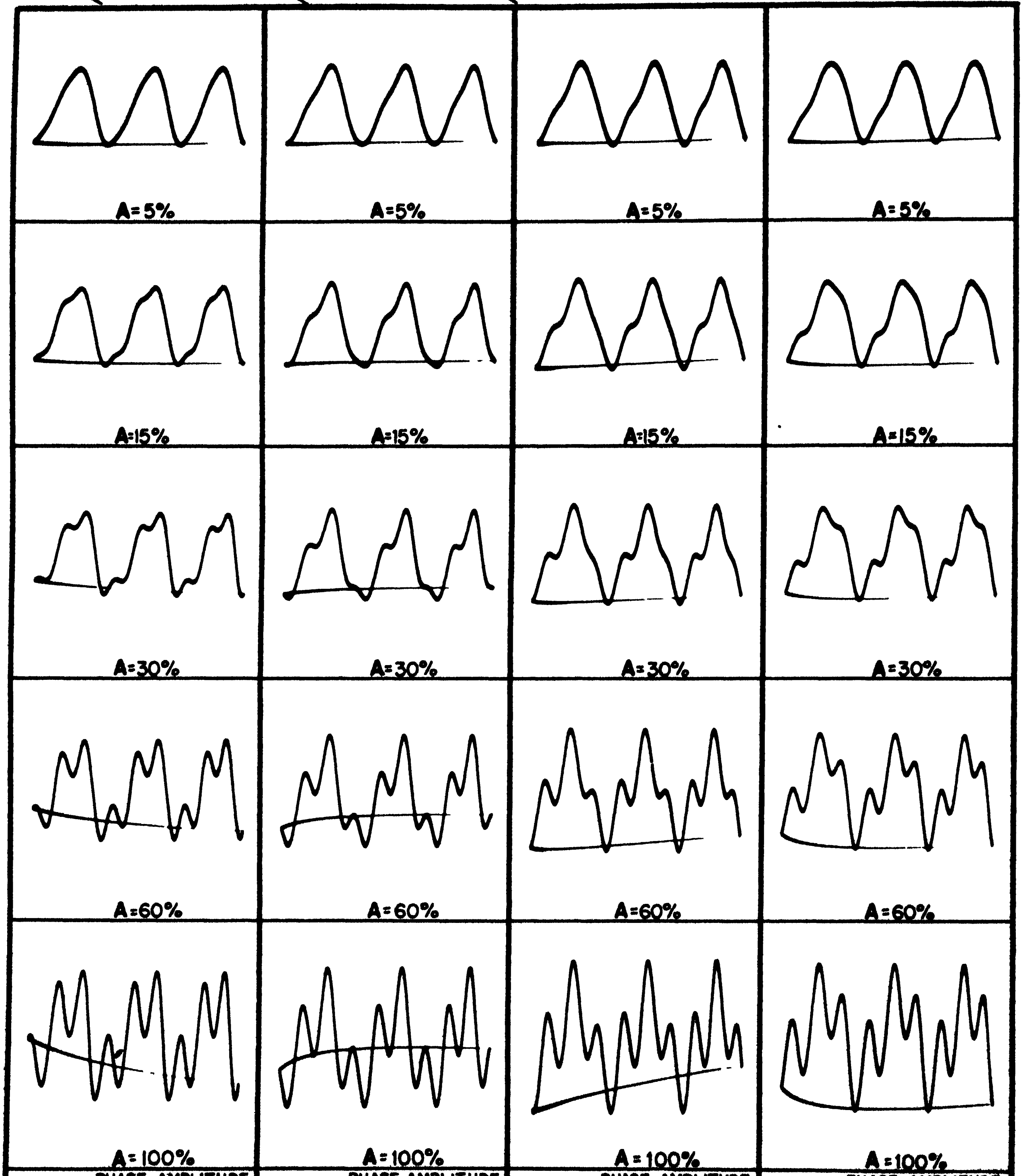
	PHASE	AMPLITUDE
Fund.	0°	100%
2nd	180°	5%
3rd	0°	see A

	PHASE	AMPLITUDE
Fund.	0°	100%
2nd	180°	5%
3rd	+90°	see A

	PHASE	AMPLITUDE
Fund.	0°	100%
2nd	180°	5%
3rd	180°	see A

	PHASE	AMPLITUDE
Fund.	0°	100%
2nd	180°	5%
3rd	-90°	see A

Fund. { 0° 100% + 2nd { 180° 15% + 3rd { 0° to −90° 5% to 100%



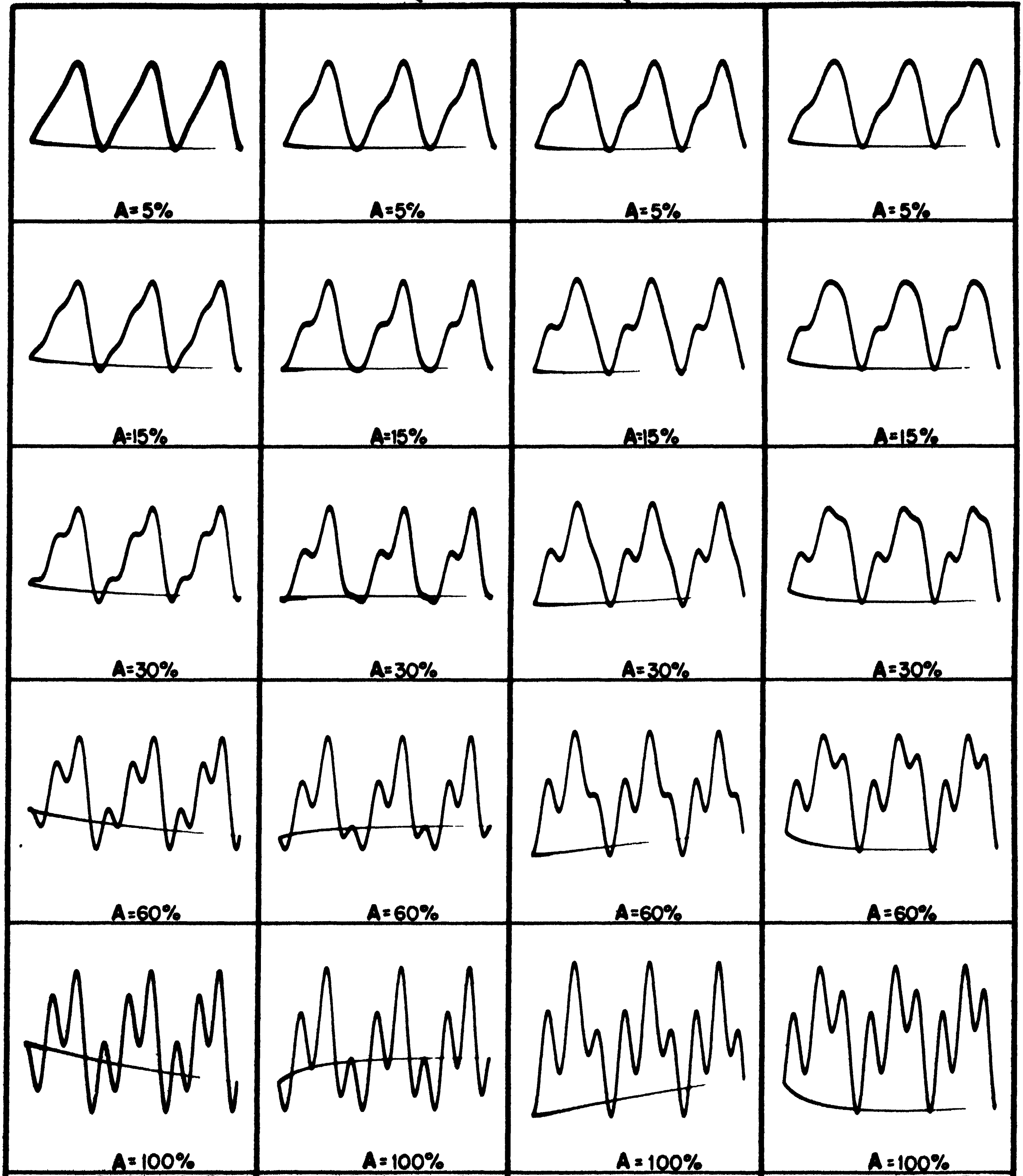
	PHASE	AMPLITUDE
Fund.	0°	100%
2nd	180°	15%
3rd	0°	see A

	PHASE	AMPLITUDE
Fund.	0°	100%
2nd	180°	15%
3rd	+90°	see A

	PHASE	AMPLITUDE
Fund.	0°	100%
2nd	180°	15%
3rd	180°	see A

	PHASE	AMPLITUDE
Fund.	0°	100%
2nd	180°	15%
3rd	−90°	see A

Fund. $\left\{\begin{matrix}0^\circ \\ 100\%\end{matrix}\right.$ + 2nd $\left\{\begin{matrix}180^\circ \\ 30\%\end{matrix}\right.$ + 3rd $\left\{\begin{matrix}0^\circ & \text{to} & -90^\circ \\ 5\% & \text{to} & 100\%\end{matrix}\right.$



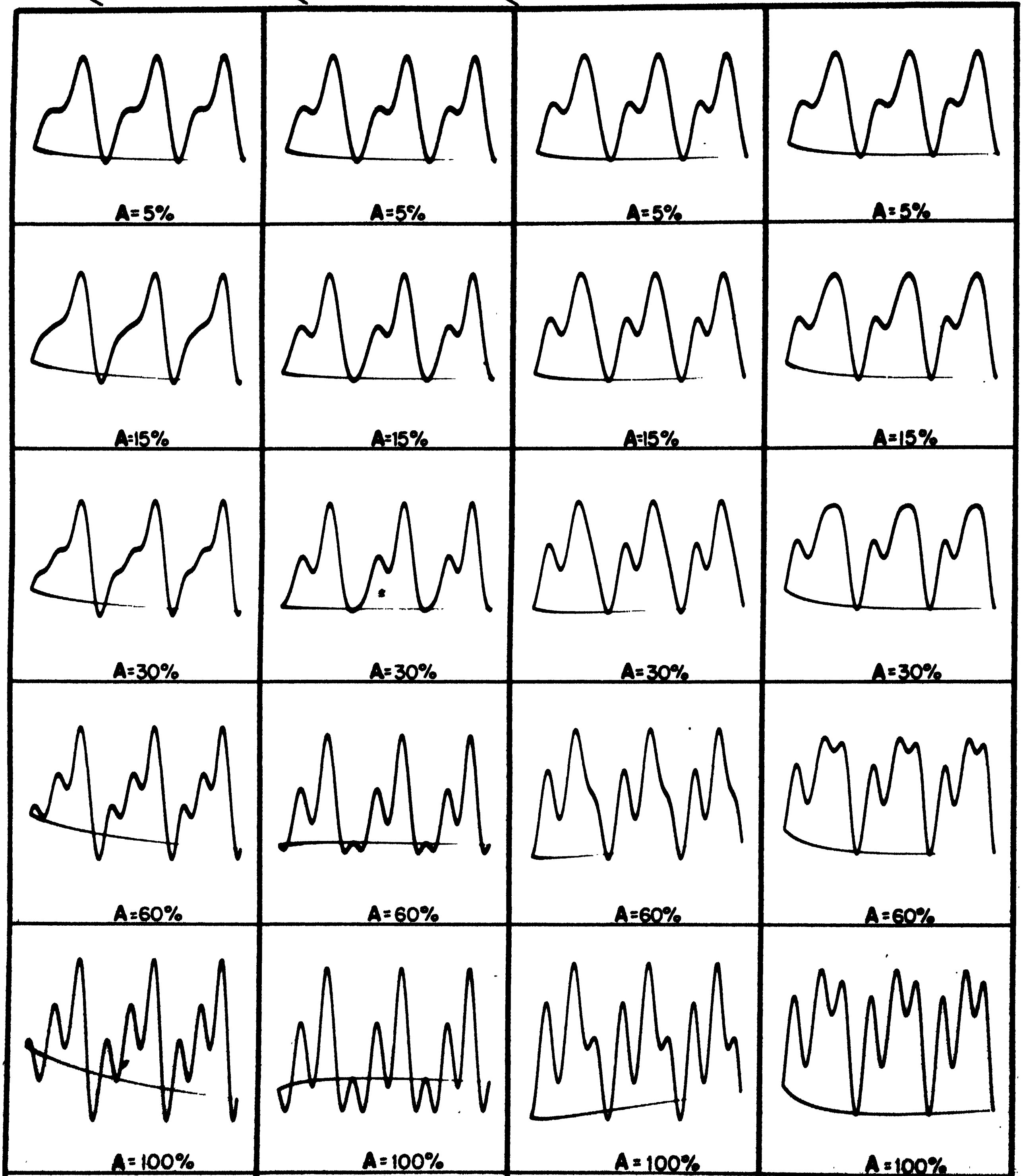
	PHASE	AMPLITUDE
Fund.	0°	100%
2nd	180°	30%
3rd	0°	*see* A

	PHASE	AMPLITUDE
Fund.	0°	100%
2nd	180°	30%
3rd	+90°	*see* A

	PHASE	AMPLITUDE
Fund.	0°	100%
2nd	180°	30%
3rd	180°	*see* A

	PHASE	AMPLITUDE
Fund.	0°	100%
2nd	180°	30%
3rd	-90°	*see* A

Fund. {0° 100% + **2nd** {180° 60% + **3rd** {0° to −90° 5% to 100%



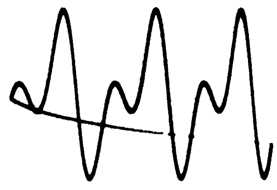
	PHASE	AMPLITUDE
Fund.	0°	100%
2nd	180°	60%
3rd	0°	*see* A

	PHASE	AMPLITUDE
Fund.	0°	100%
2nd	180°	60%
3rd	+90°	*see* A

	PHASE	AMPLITUDE
Fund.	0°	100%
2nd	180°	60%
3rd	180°	*see* A

	PHASE	AMPLITUDE
Fund.	0°	100%
2nd	180°	60%
3rd	−90°	*see* A

$$\text{Fund.}\left\{\begin{matrix}0^\circ\\100\%\end{matrix}\right. + \text{2nd}\left\{\begin{matrix}180^\circ\\100\%\end{matrix}\right. + \text{3rd}\left\{\begin{matrix}0^\circ & \text{to} & -90^\circ\\5\% & \text{to} & 100\%\end{matrix}\right.$$



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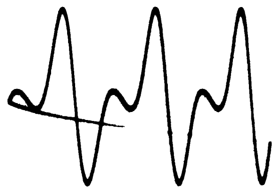

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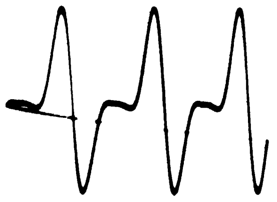

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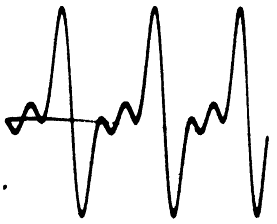

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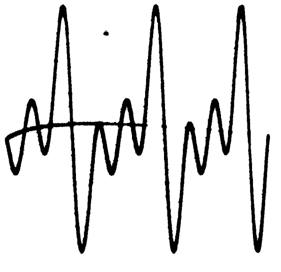
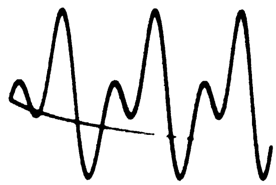


A=100%

	PHASE	AMPLITUDE
Fund.	0°	100%
2nd	180°	100%
3rd	0°	*see A*



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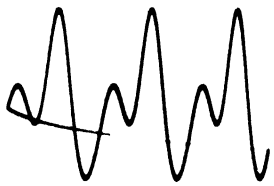

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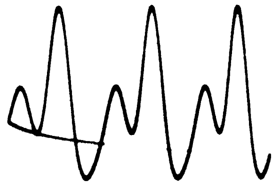

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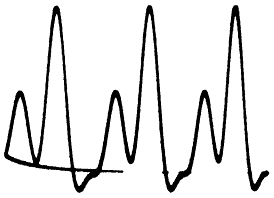

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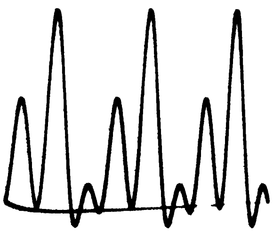
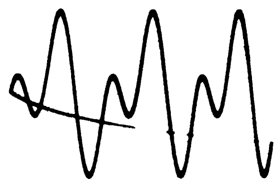


A=100%

	PHASE	AMPLITUDE
Fund.	0°	100%
2nd	180°	100%
3rd	+90°	*see A*



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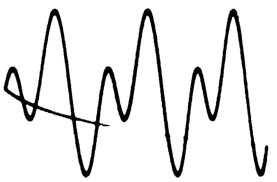

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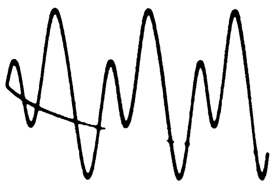

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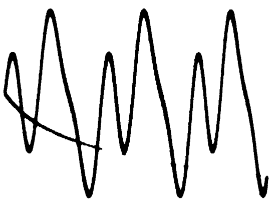

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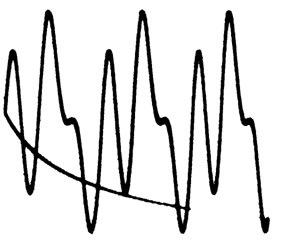

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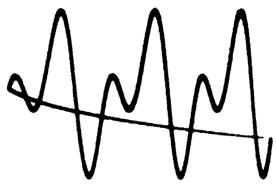
	PHASE	AMPLITUDE
Fund.	0°	100%
2nd	180°	100%
3rd	180°	*see A*



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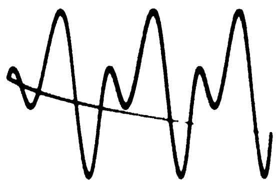

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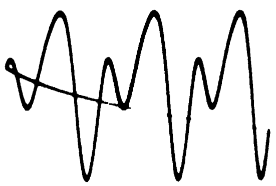

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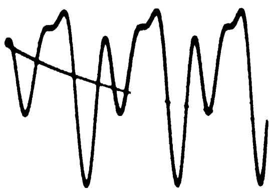

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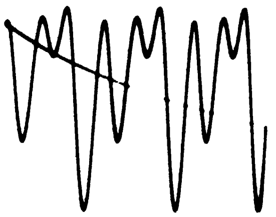
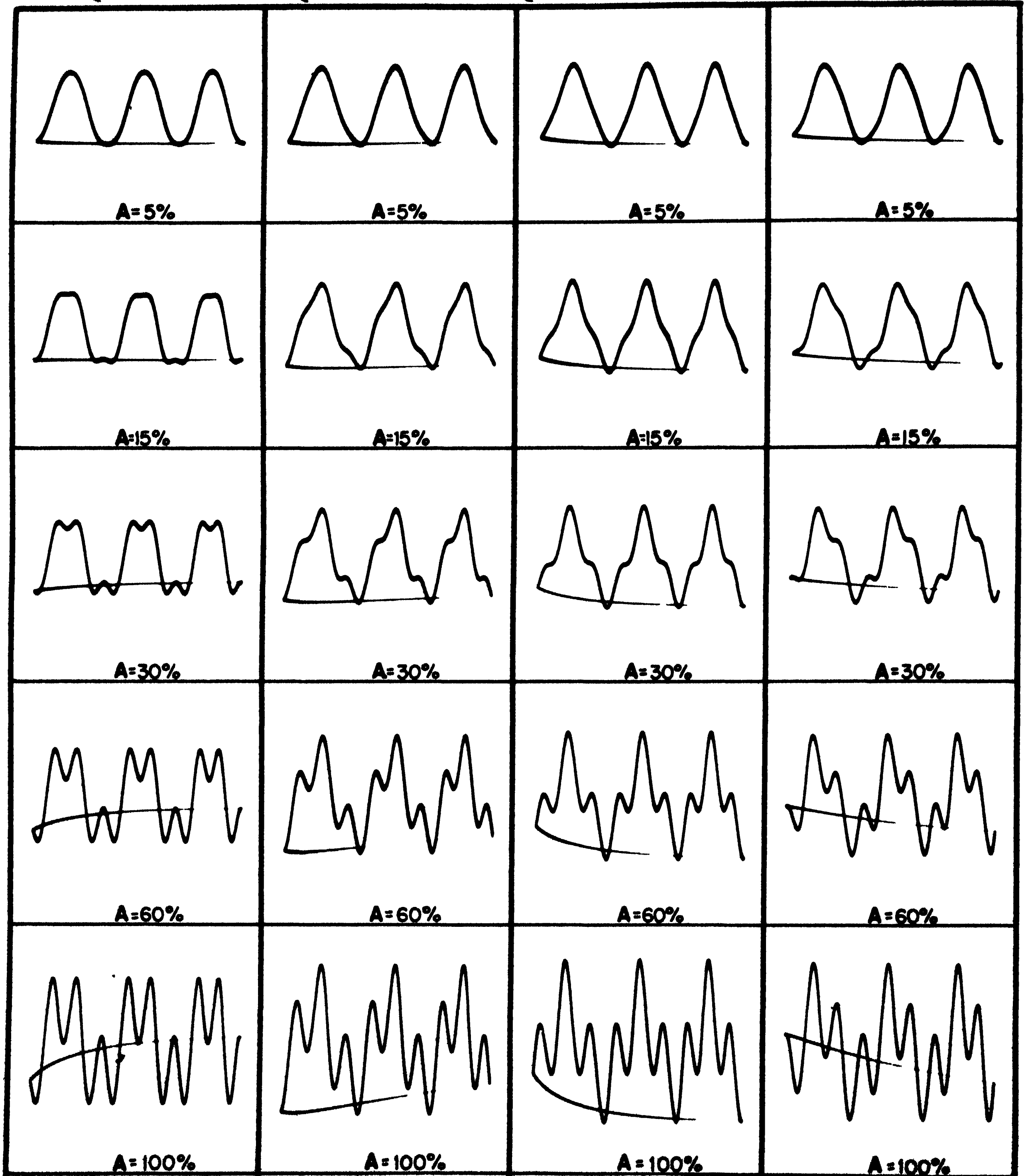


A=100%

	PHASE	AMPLITUDE
Fund.	0°	100%
2nd	180°	100%
3rd	-90°	*see A*

Fund. { 0° 100% + **2nd** { −90° 5% + **3rd** { 0° to −90° 5% to 100%



A=5% A=5% A=5% A=5%

A=15% A=15% A=15% A=15%

A=30% A=30% A=30% A=30%

A=60% A=60% A=60% A=60%

A=100% A=100% A=100% A=100%

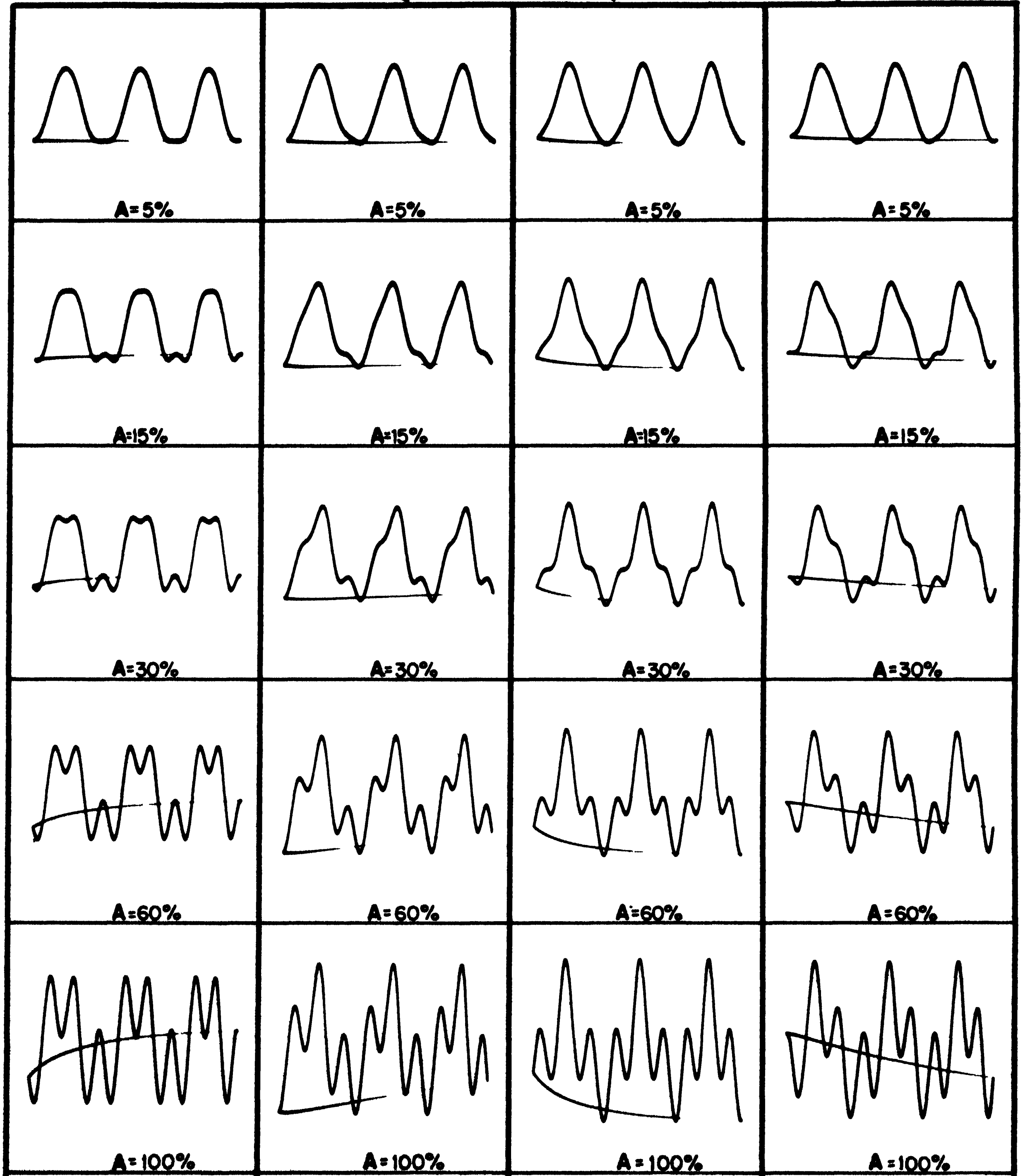
	PHASE	AMPLITUDE
Fund.	0°	100%
2nd	-90°	5%
3rd	0°	see A

	PHASE	AMPLITUDE
Fund.	0°	100%
2nd	-90°	5%
3rd	+90°	see A

	PHASE	AMPLITUDE
Fund.	0°	100%
2nd	-90°	5%
3rd	180°	see A

	PHASE	AMPLITUDE
Fund.	0°	100%
2nd	-90°	5%
3rd	-90°	see A

$$\text{Fund.}\begin{cases}0^\circ\\100\%\end{cases} + \text{2nd}\begin{cases}-90^\circ\\15\%\end{cases} + \text{3rd}\begin{cases}0^\circ \text{ to } -90^\circ\\5\% \text{ to } 100\%\end{cases}$$



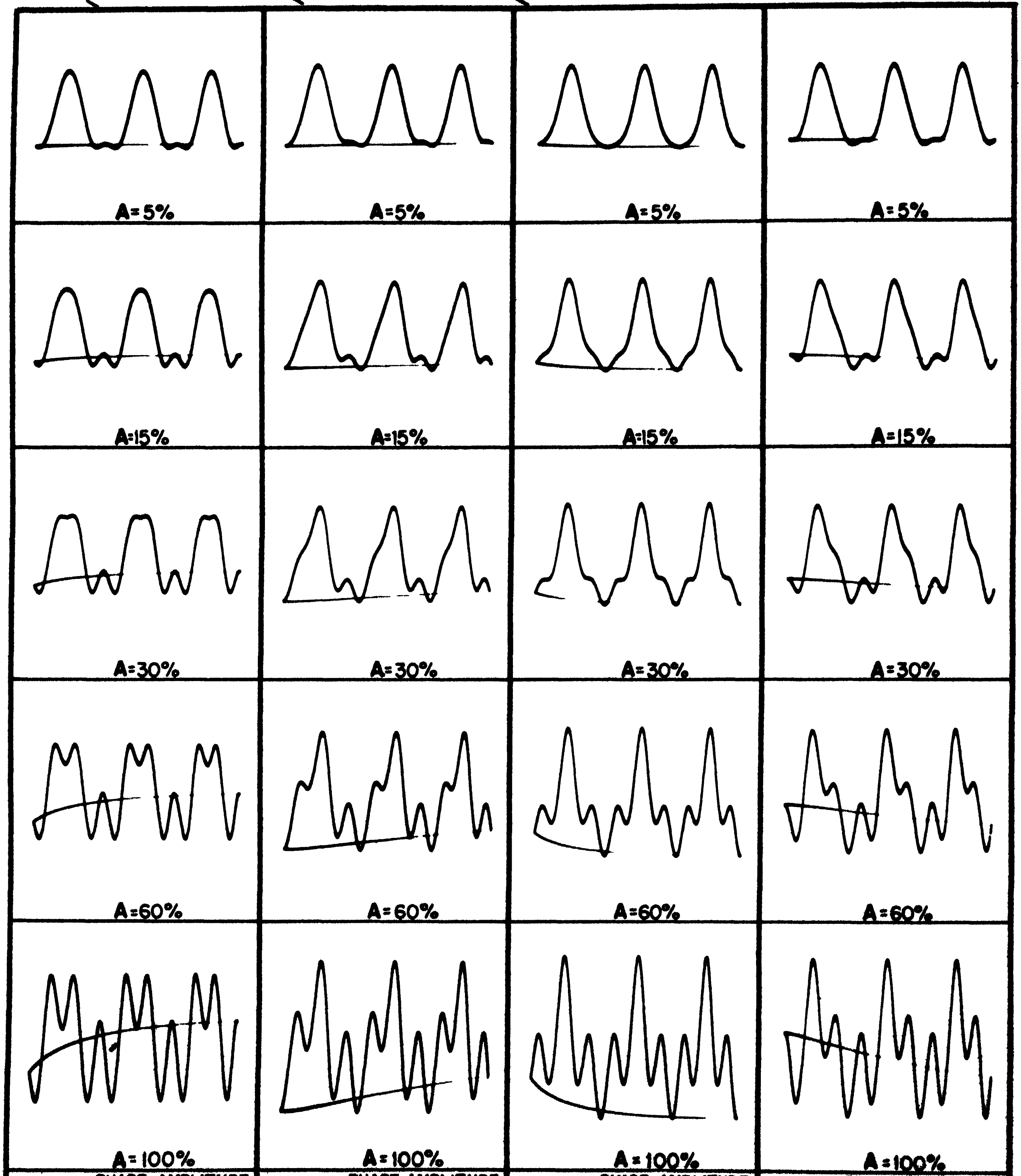
	PHASE	AMPLITUDE
Fund.	0°	100%
2nd	-90°	15%
3rd	0°	see A

	PHASE	AMPLITUDE
Fund.	0°	100%
2nd	-90°	15%
3rd	+90°	see A

	PHASE	AMPLITUDE
Fund.	0°	100%
2nd	-90°	15%
3rd	180°	see A

	PHASE	AMPLITUDE
Fund.	0°	100%
2nd	-90°	15%
3rd	-90°	see A

Fund. $\left\{\begin{matrix}0^\circ\\100\%\end{matrix}\right.$ **+ 2nd** $\left\{\begin{matrix}-90^\circ\\30\%\end{matrix}\right.$ **+ 3rd** $\left\{\begin{matrix}0^\circ \text{ to } -90^\circ\\5\% \text{ to } 100\%\end{matrix}\right.$



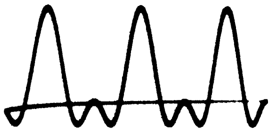
	PHASE	AMPLITUDE
Fund.	0°	100%
2nd	-90°	30%
3rd	0°	see A

	PHASE	AMPLITUDE
Fund.	0°	100%
2nd	-90°	30%
3rd	+90°	see A

	PHASE	AMPLITUDE
Fund.	0°	100%
2nd	-90°	30%
3rd	180°	see A

	PHASE	AMPLITUDE
Fund.	0°	100%
2nd	-90°	30%
3rd	-90°	see A

$$\text{Fund.}\begin{cases}0^\circ\\100\%\end{cases}+\ \text{2nd}\begin{cases}-90^\circ\\60\%\end{cases}+\ \text{3rd}\begin{cases}0^\circ\ \text{to}\ -90^\circ\\5\%\ \text{to}\ 100\%\end{cases}$$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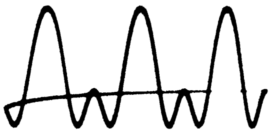
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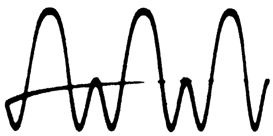
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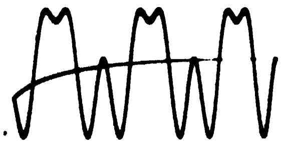
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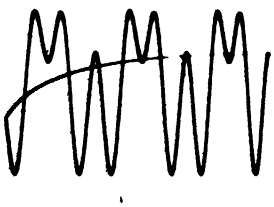
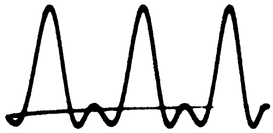

A=100%

	PHASE	AMPLITUDE
Fund.	0°	100%
2nd	-90°	60%
3rd	0°	*see A*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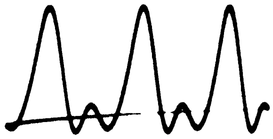
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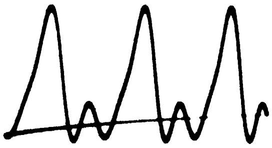
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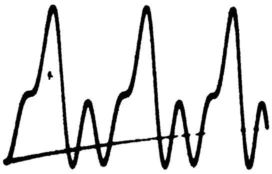
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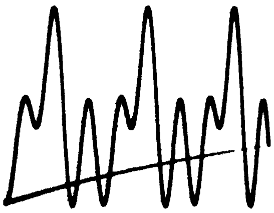
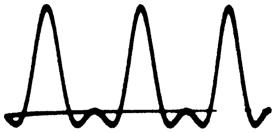

A=100%

	PHASE	AMPLITUDE
Fund.	0°	100%
2nd	-90°	60%
3rd	+90°	*see A*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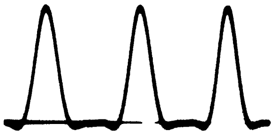
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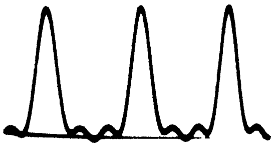
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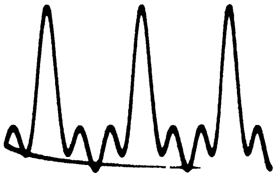
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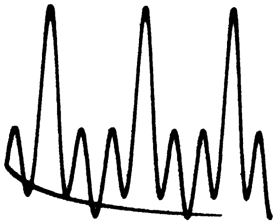
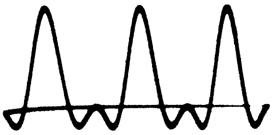

A=100%

	PHASE	AMPLITUDE
Fund.	0°	100%
2nd	-90°	60%
3rd	180°	*see A*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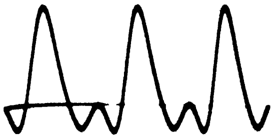
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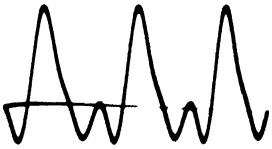
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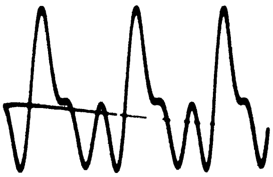
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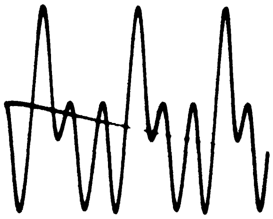
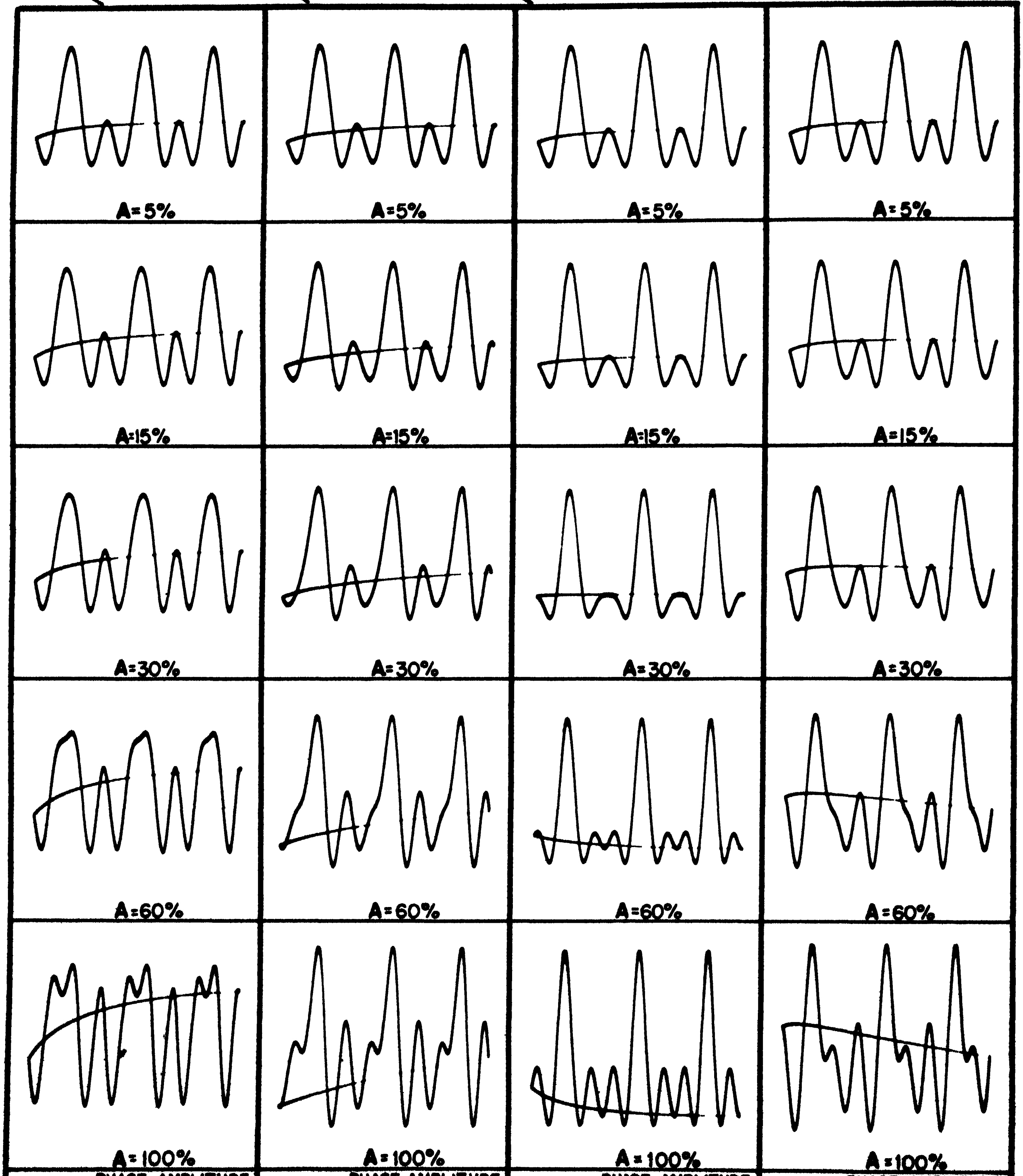

A=100%

	PHASE	AMPLITUDE
Fund.	0°	100%
2nd	-90°	60%
3rd	-90°	*see A*

$$\text{Fund.}\begin{cases}0^\circ\\100\%\end{cases} + \text{2nd}\begin{cases}-90^\circ\\100\%\end{cases} + \text{3rd}\begin{cases}0^\circ \text{ to } -90^\circ\\5\% \text{ to } 100\%\end{cases}$$



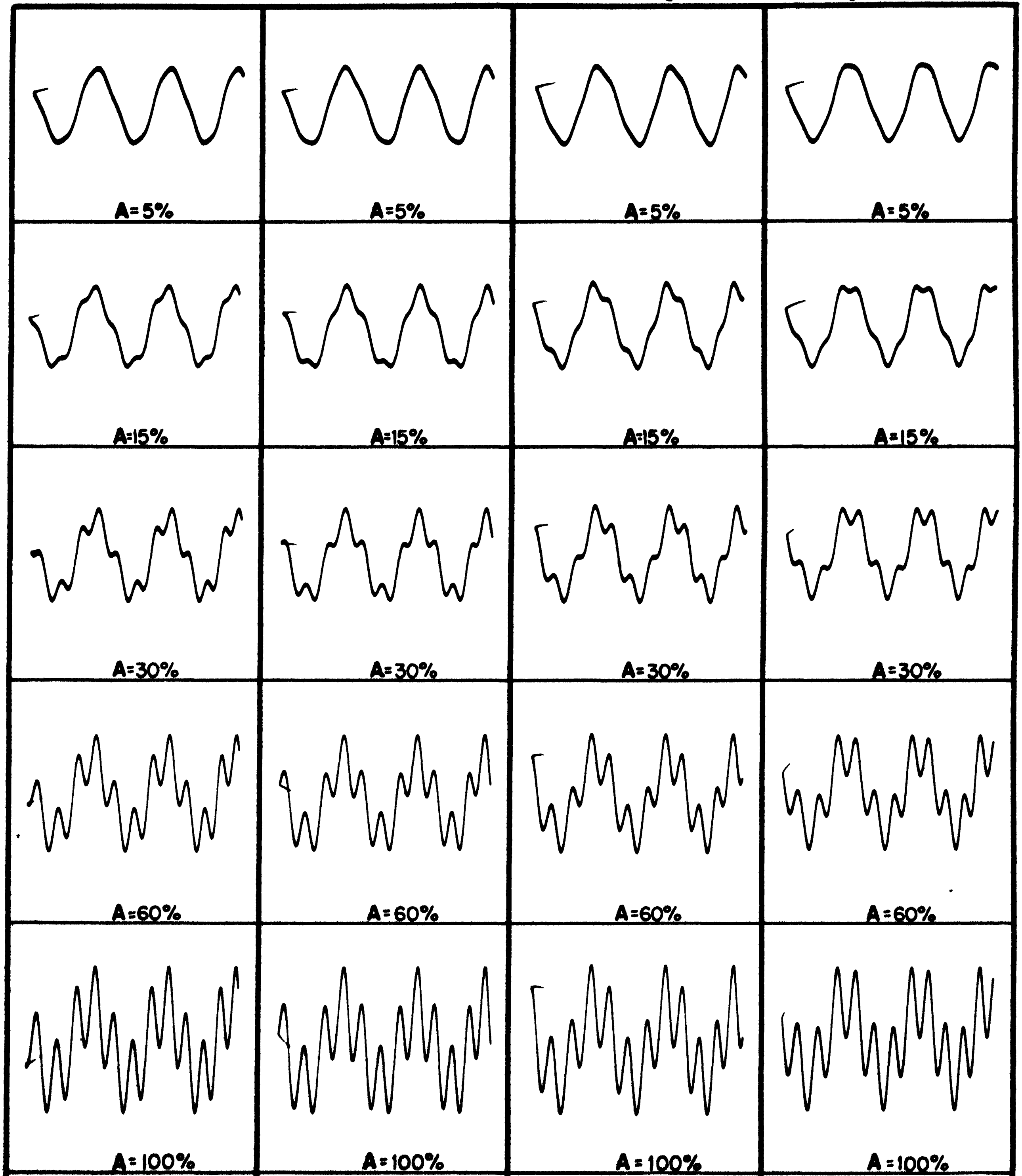
	PHASE	AMPLITUDE
Fund.	0°	100%
2nd	-90°	100%
3rd	0°	see A

	PHASE	AMPLITUDE
Fund.	0°	100%
2nd	-90°	100%
3rd	+90°	see A

	PHASE	AMPLITUDE
Fund.	0°	100%
2nd	-90°	100%
3rd	180°	see A

	PHASE	AMPLITUDE
Fund.	0°	100%
2nd	-90°	100%
3rd	-90°	see A

$$\text{Fund.}\left\{\begin{matrix}0^\circ\\100\%\end{matrix}\right. + \text{2nd}\left\{\begin{matrix}0^\circ\\5\%\end{matrix}\right. + \text{4th}\left\{\begin{matrix}0^\circ \text{ to } -90^\circ\\5\% \text{ to } 100\%\end{matrix}\right.$$



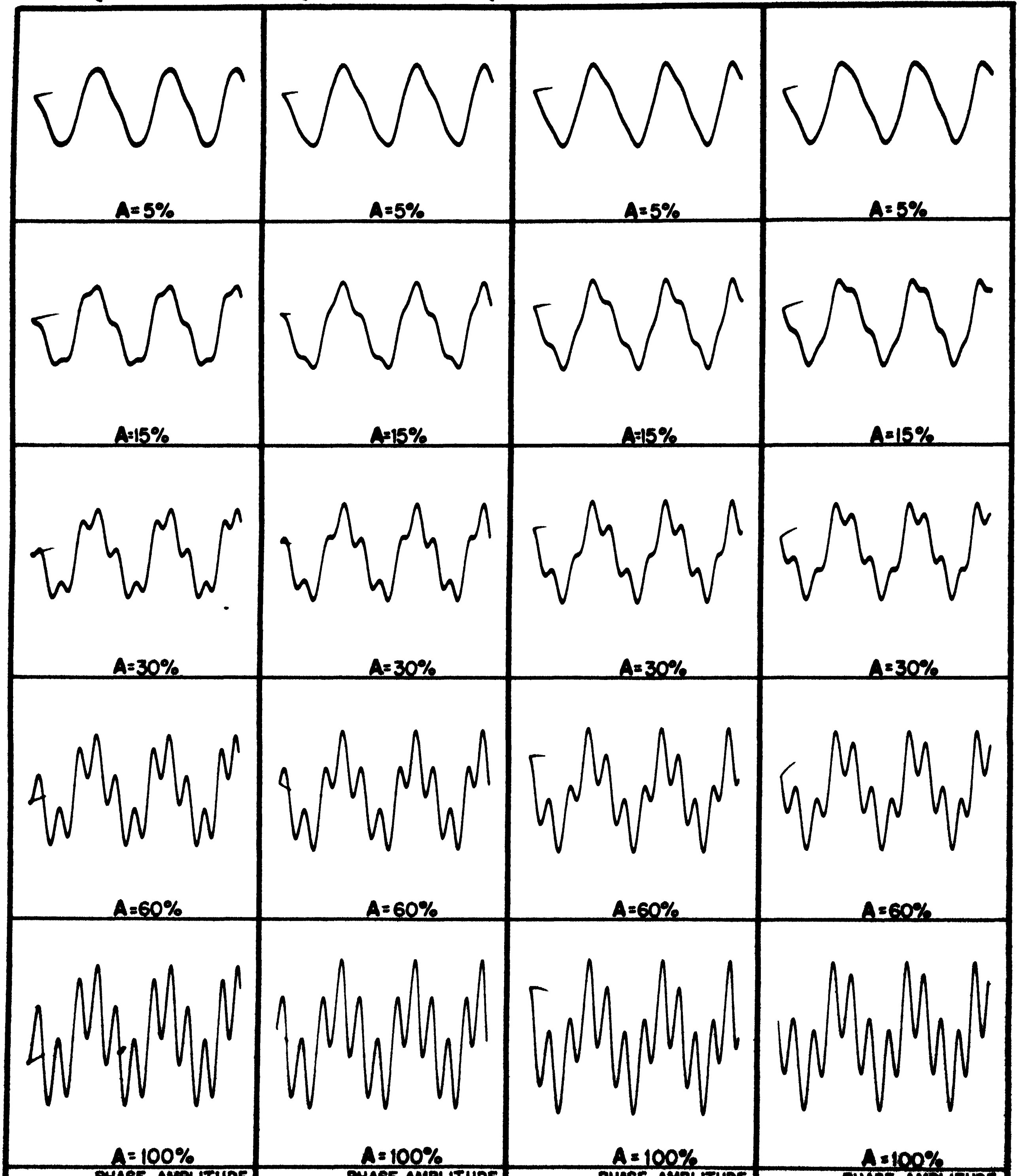
	PHASE	AMPLITUDE
Fund.	0°	100%
2nd	0°	5%
4th	0°	*see* A

	PHASE	AMPLITUDE
Fund.	0°	100%
2nd	0°	5%
4th	+90°	*see* A

	PHASE	AMPLITUDE
Fund.	0°	100%
2nd	0°	5%
4th	180°	*see* A

	PHASE	AMPLITUDE
Fund.	0°	100%
2nd	0°	5%
4th	-90°	*see* A

$$\text{Fund.}\begin{cases}0^\circ\\100\%\end{cases} + \text{2nd}\begin{cases}0^\circ\\15\%\end{cases} + \text{4th}\begin{cases}0^\circ \text{ to } -90^\circ\\5\% \text{ to } 100\%\end{cases}$$



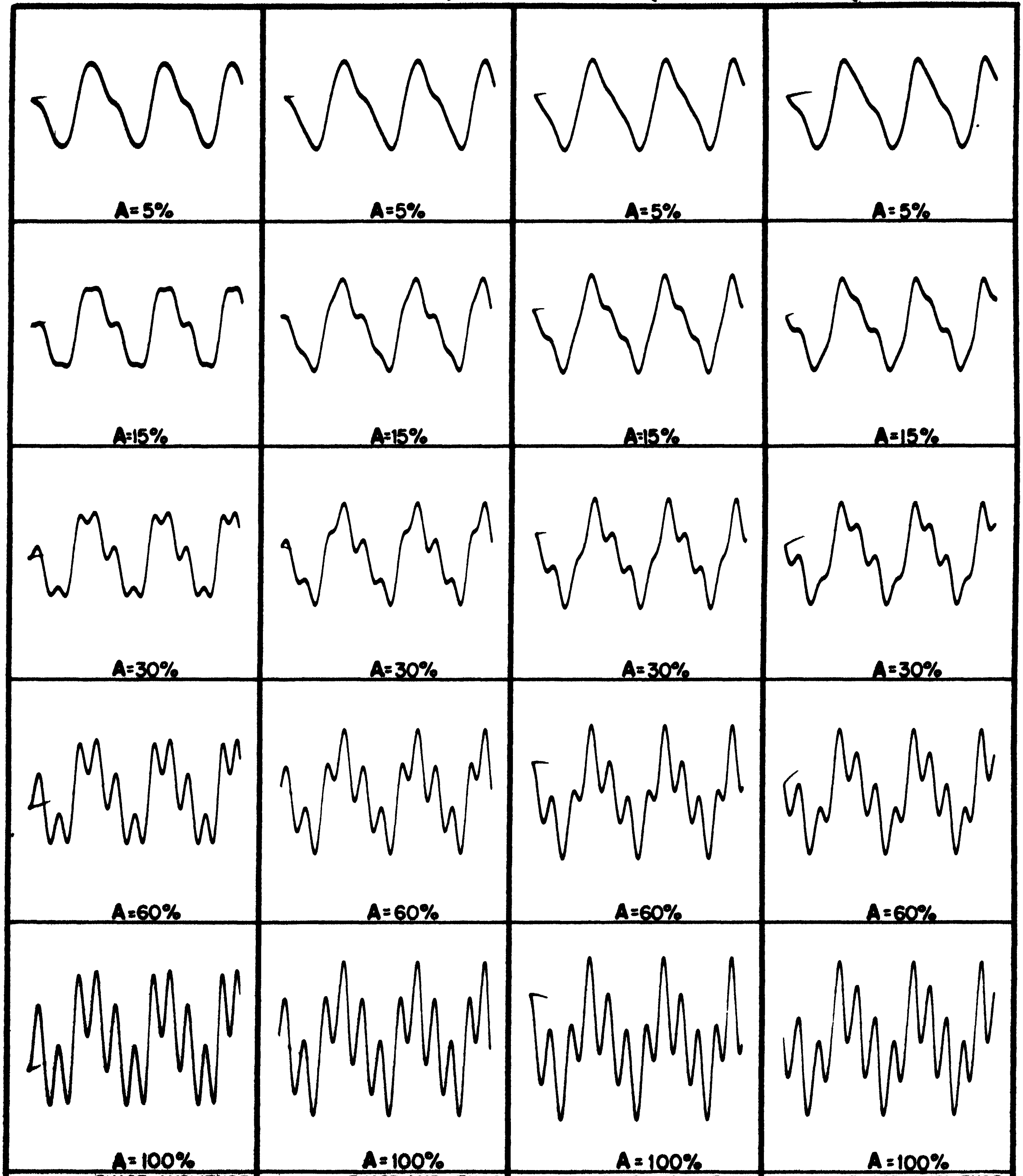
	PHASE	AMPLITUDE
Fund.	0°	100%
2nd	0°	15%
4th	0°	see A

	PHASE	AMPLITUDE
Fund.	0°	100%
2nd	0°	15%
4th	+90°	see A

	PHASE	AMPLITUDE
Fund.	0°	100%
2nd	0°	15%
4th	180°	see A

	PHASE	AMPLITUDE
Fund.	0°	100%
2nd	0°	15%
4th	-90°	see A

$$\text{Fund.}\left\{\begin{matrix}0^\circ\\100\%\end{matrix}\right. + \text{2nd}\left\{\begin{matrix}0^\circ\\30\%\end{matrix}\right. + \text{4th}\left\{\begin{matrix}0^\circ \text{ to } -90^\circ\\5\% \text{ to } 100\%\end{matrix}\right.$$



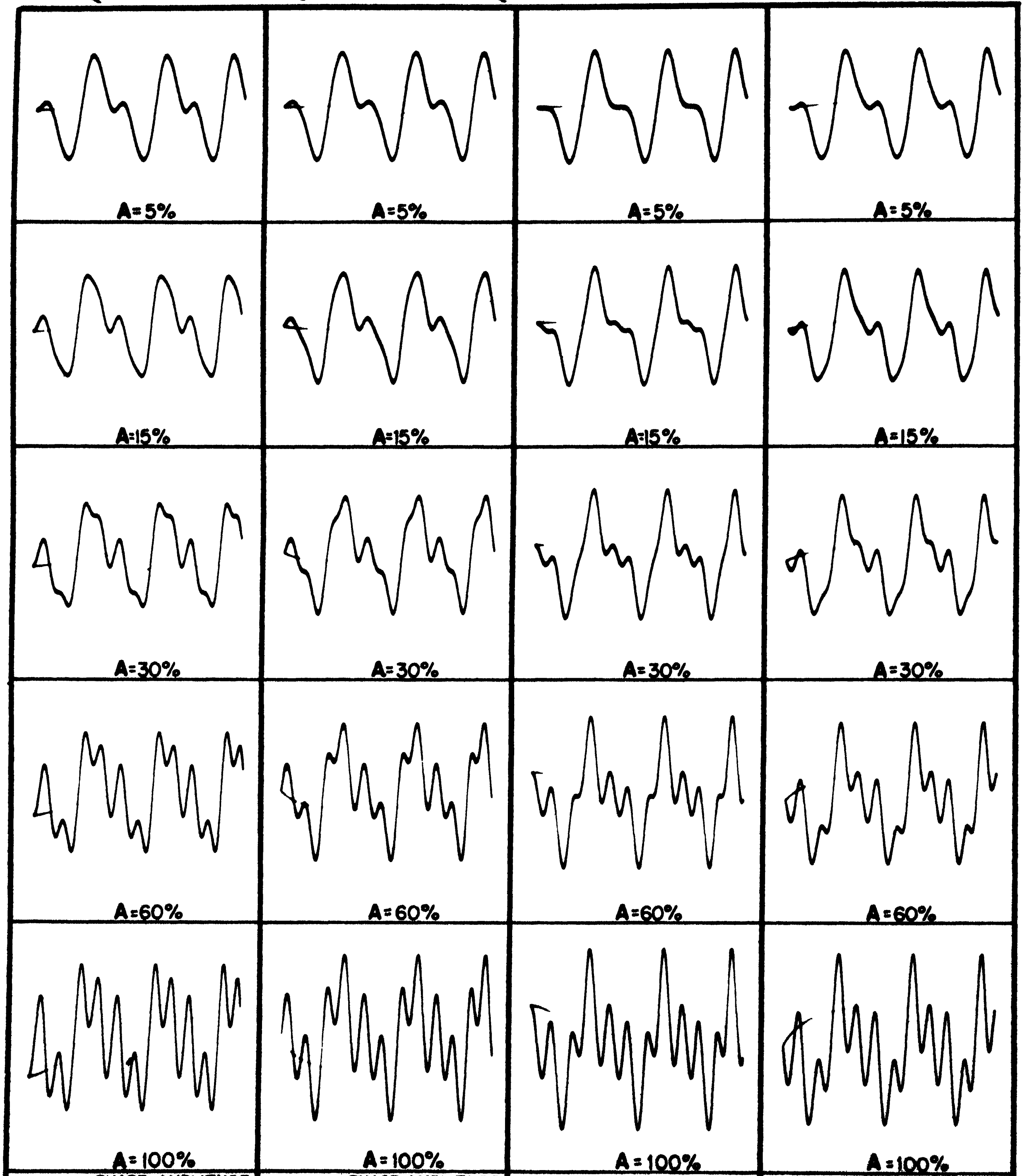
	PHASE	AMPLITUDE
Fund.	0°	100%
2nd	0°	30%
4th	0°	*see* A

	PHASE	AMPLITUDE
Fund.	0°	100%
2nd	0°	30%
4th	+90°	*see* A

	PHASE	AMPLITUDE
Fund.	0°	100%
2nd	0°	30%
4th	180°	*see* A

	PHASE	AMPLITUDE
Fund.	0°	100%
2nd	0°	30%
4th	-90°	*see* A

$$\text{Fund.}\begin{cases}0^\circ\\100\%\end{cases} + \text{2nd}\begin{cases}0^\circ\\60\%\end{cases} + \text{4th}\begin{cases}0^\circ \text{ to } -90^\circ\\5\% \text{ to } 100\%\end{cases}$$



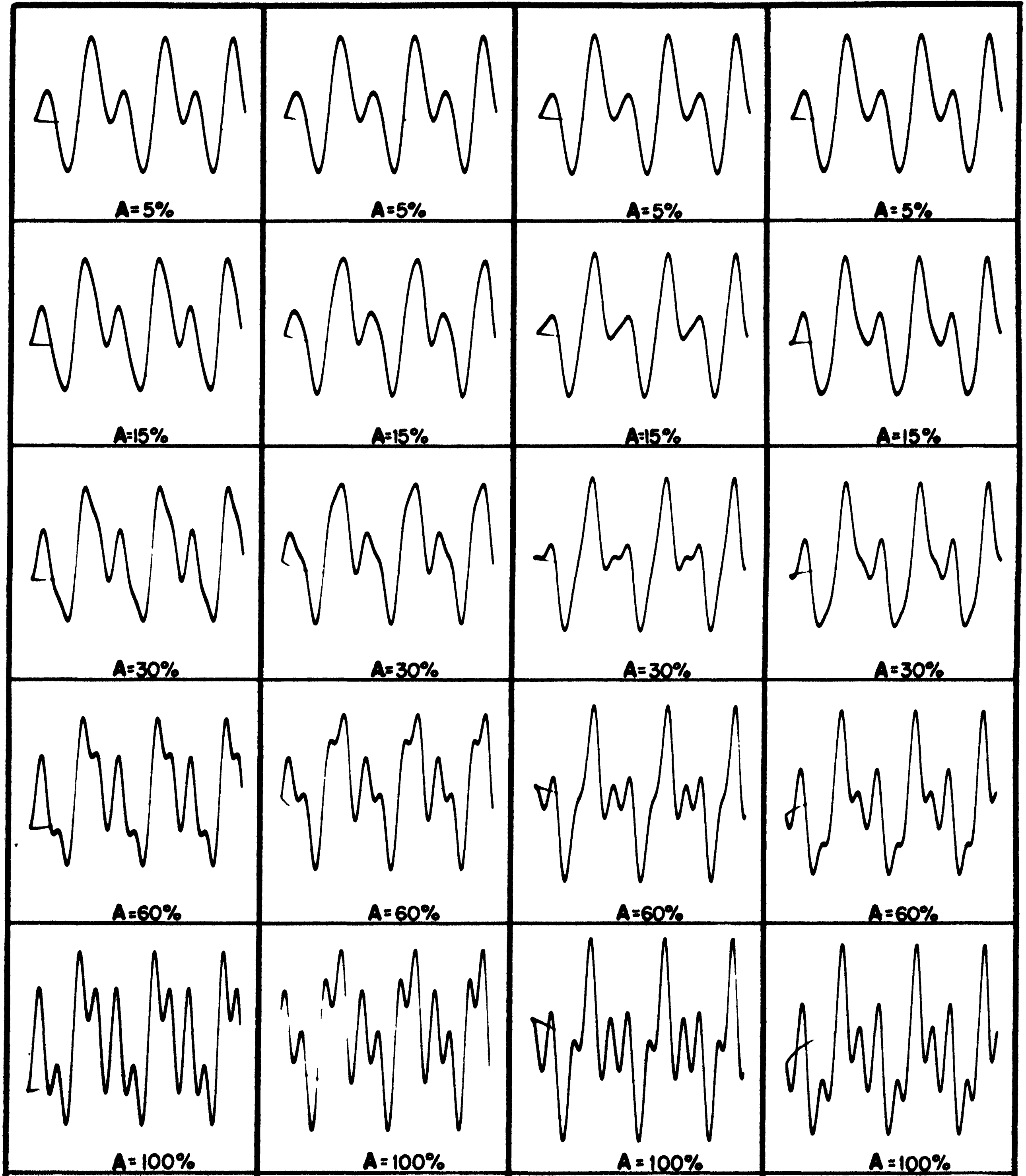
	PHASE	AMPLITUDE
Fund.	0°	100%
2nd	0°	60%
4th	0°	*see* A

	PHASE	AMPLITUDE
Fund.	0°	100%
2nd	0°	60%
4th	+90°	*see* A

	PHASE	AMPLITUDE
Fund.	0°	100%
2nd	0°	60%
4th	180°	*see* A

	PHASE	AMPLITUDE
Fund.	0°	100%
2nd	0°	60%
4th	-90°	*see* A

Fund. $\begin{cases} 0° \\ 100\% \end{cases}$ + 2nd $\begin{cases} 0° \\ 100\% \end{cases}$ + 4th $\begin{cases} 0° \text{ to } -90° \\ 5\% \text{ to } 100\% \end{cases}$



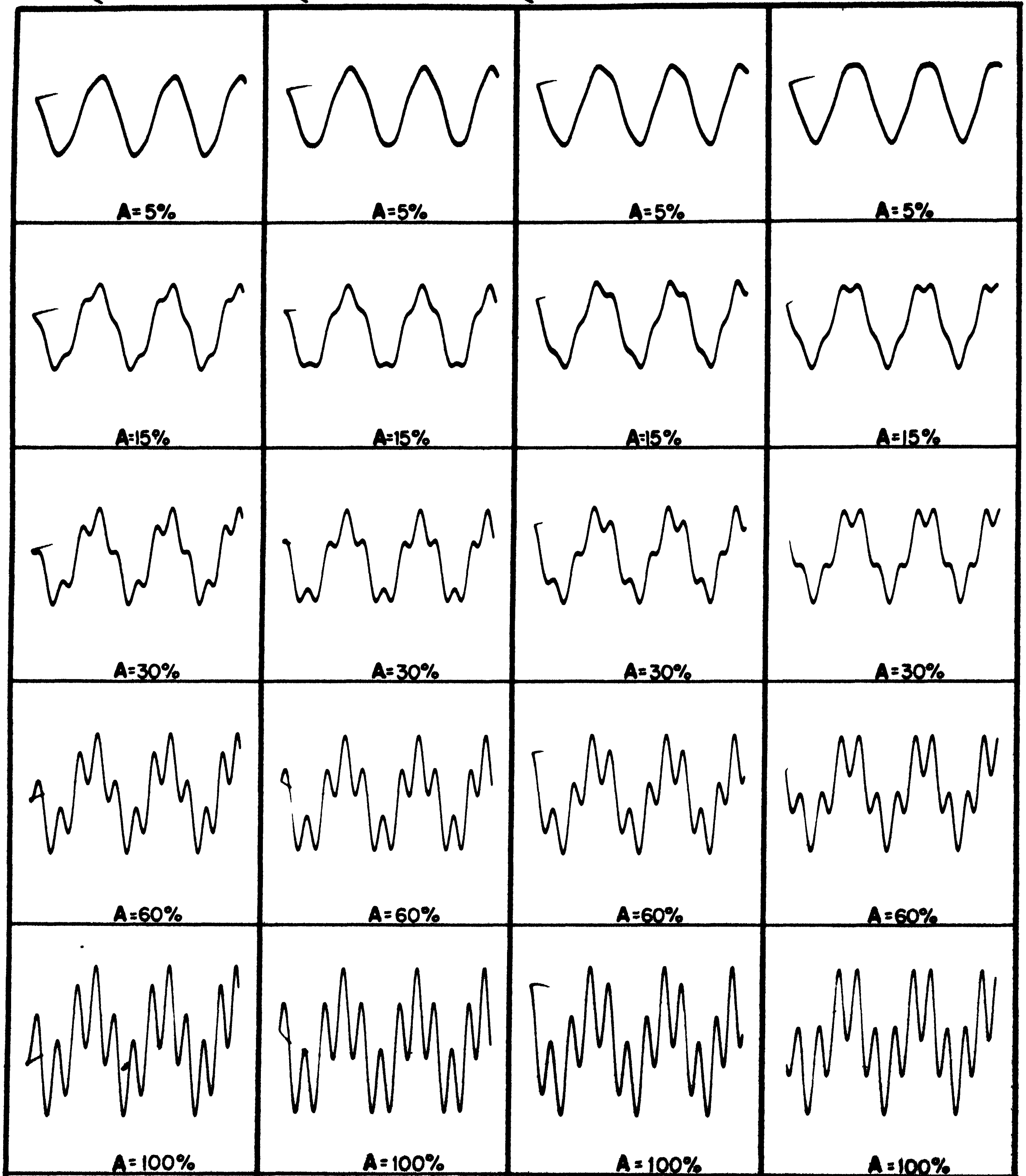
	PHASE	AMPLITUDE
Fund.	0°	100%
2nd	0°	100%
4th	0°	*see* A

	PHASE	AMPLITUDE
Fund.	0°	100%
2nd	0°	100%
4th	+90°	*see* A

	PHASE	AMPLITUDE
Fund.	0°	100%
2nd	0°	100%
4th	180°	*see* A

	PHASE	AMPLITUDE
Fund.	0°	100%
2nd	0°	100%
4th	-90°	*see* A

$$\text{Fund.}\left\{\begin{matrix}0^\circ\\100\%\end{matrix}\right. + \text{2nd}\left\{\begin{matrix}+90^\circ\\5\%\end{matrix}\right. + \text{4th}\left\{\begin{matrix}0^\circ \text{ to } -90^\circ\\5\% \text{ to } 100\%\end{matrix}\right.$$



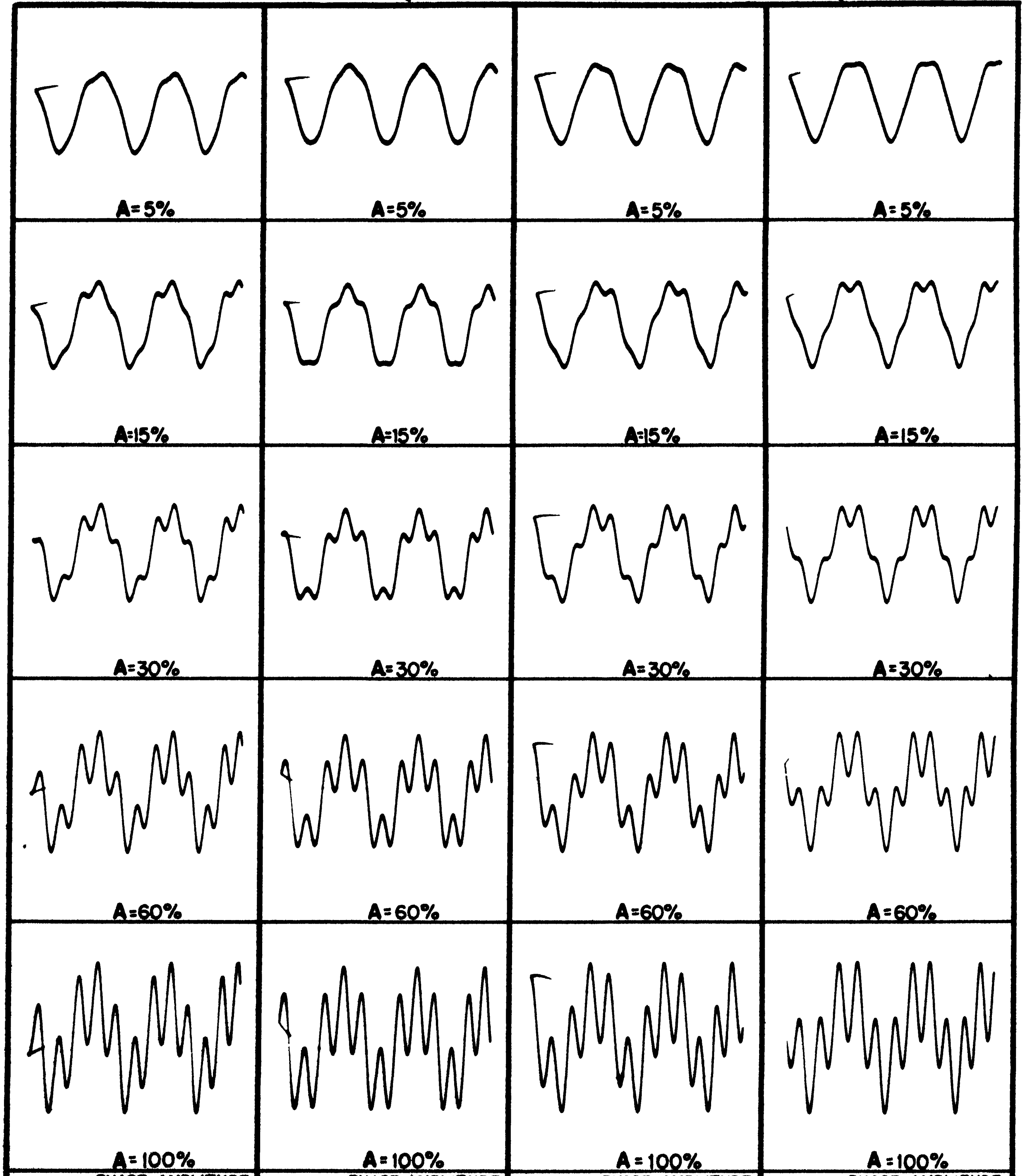
	PHASE	AMPLITUDE
Fund.	0°	100%
2nd	+90°	5%
4th	0°	*see* A

	PHASE	AMPLITUDE
Fund.	0°	100%
2nd	+90°	5%
4th	+90°	*see* A

	PHASE	AMPLITUDE
Fund.	0°	100%
2nd	+90°	5%
4th	180°	*see* A

	PHASE	AMPLITUDE
Fund.	0°	100%
2nd	+90°	5%
4th	-90°	*see* A

$$\text{Fund.}\begin{cases}0^\circ\\100\%\end{cases} + \text{2nd}\begin{cases}+90^\circ\\15\%\end{cases} + \text{4th}\begin{cases}0^\circ \text{ to } -90^\circ\\5\% \text{ to } 100\%\end{cases}$$



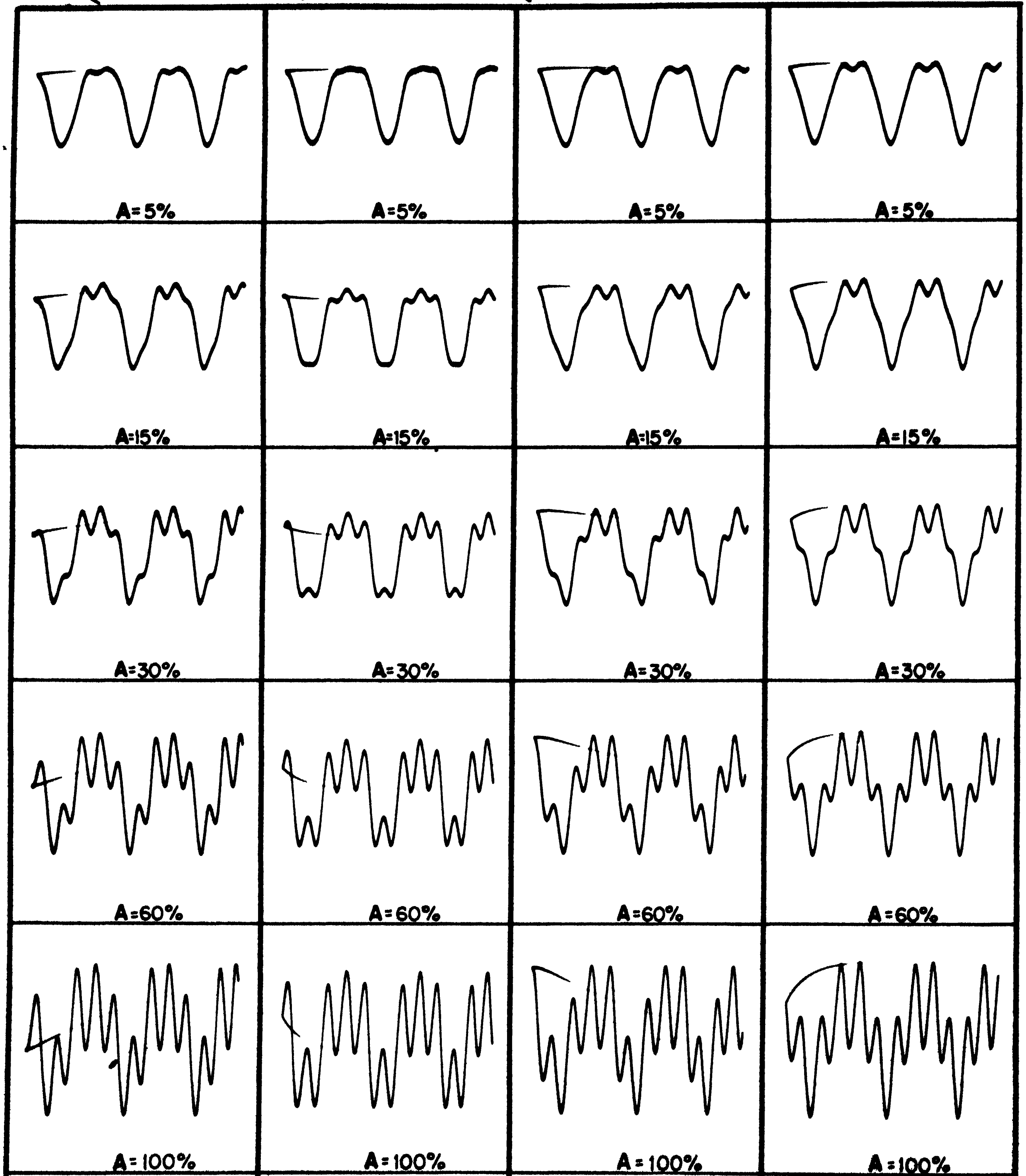
	PHASE	AMPLITUDE
Fund.	0°	100%
2nd	+90°	15%
4th	0°	*see* A

	PHASE	AMPLITUDE
Fund.	0°	100%
2nd	+90°	15%
4th	+90°	*see* A

	PHASE	AMPLITUDE
Fund.	0°	100%
2nd	+90°	15%
4th	180°	*see* A

	PHASE	AMPLITUDE
Fund.	0°	100%
2nd	+90°	15%
4th	-90°	*see* A

Fund. { 0° 100% + 2nd { +90° 30% + 4th { 0° to −90° 5% to 100%



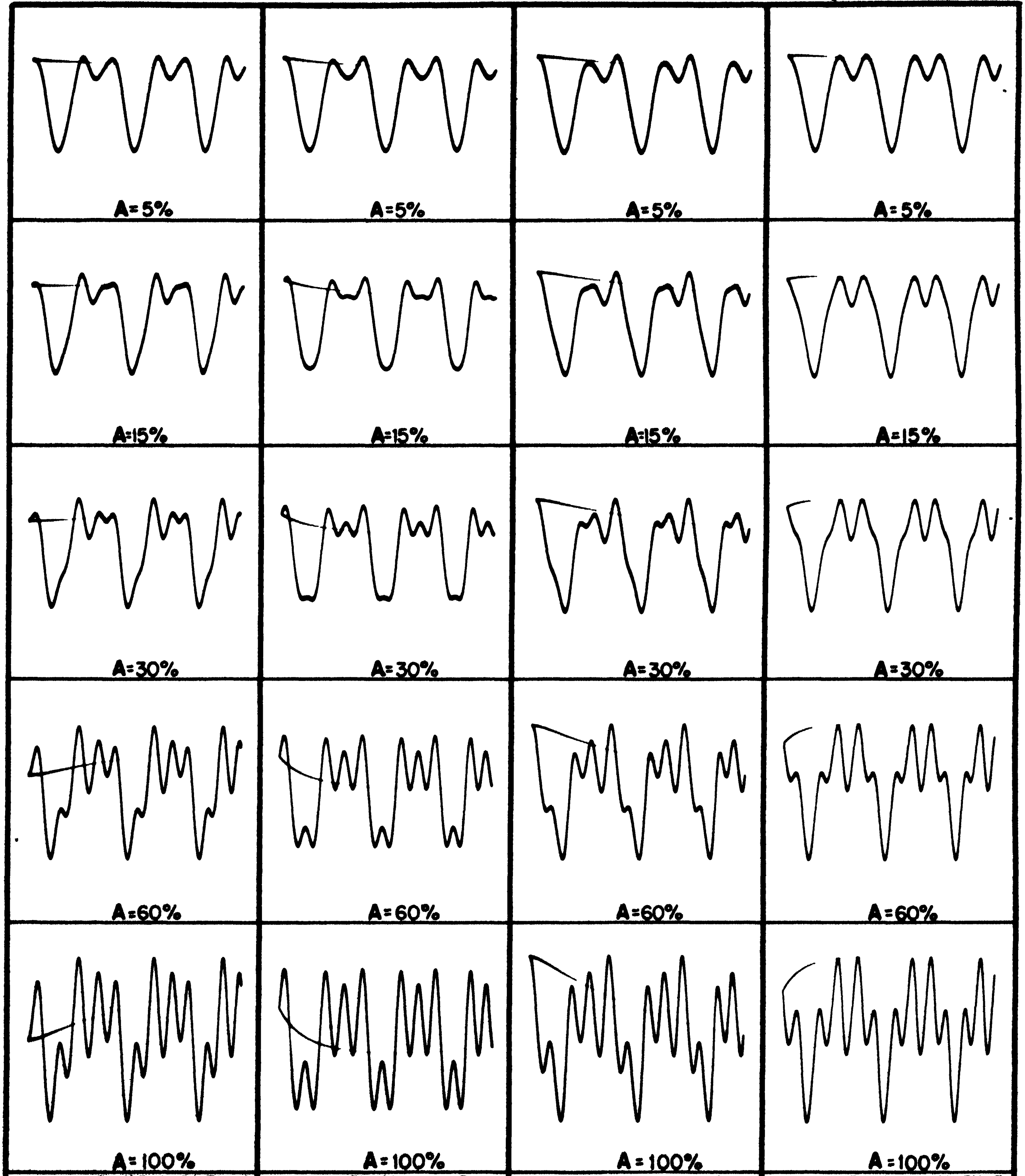
	PHASE	AMPLITUDE
Fund.	0°	100%
2nd	+90°	30%
4th	0°	see A

	PHASE	AMPLITUDE
Fund.	0°	100%
2nd	+90°	30%
4th	+90°	see A

	PHASE	AMPLITUDE
Fund.	0°	100%
2nd	+90°	30%
4th	180°	see A

	PHASE	AMPLITUDE
Fund.	0°	100%
2nd	+90°	30%
4th	−90°	see A

Fund. {0° 100% + 2nd {+90° 60% + 4th {0° to −90° 5% to 100%



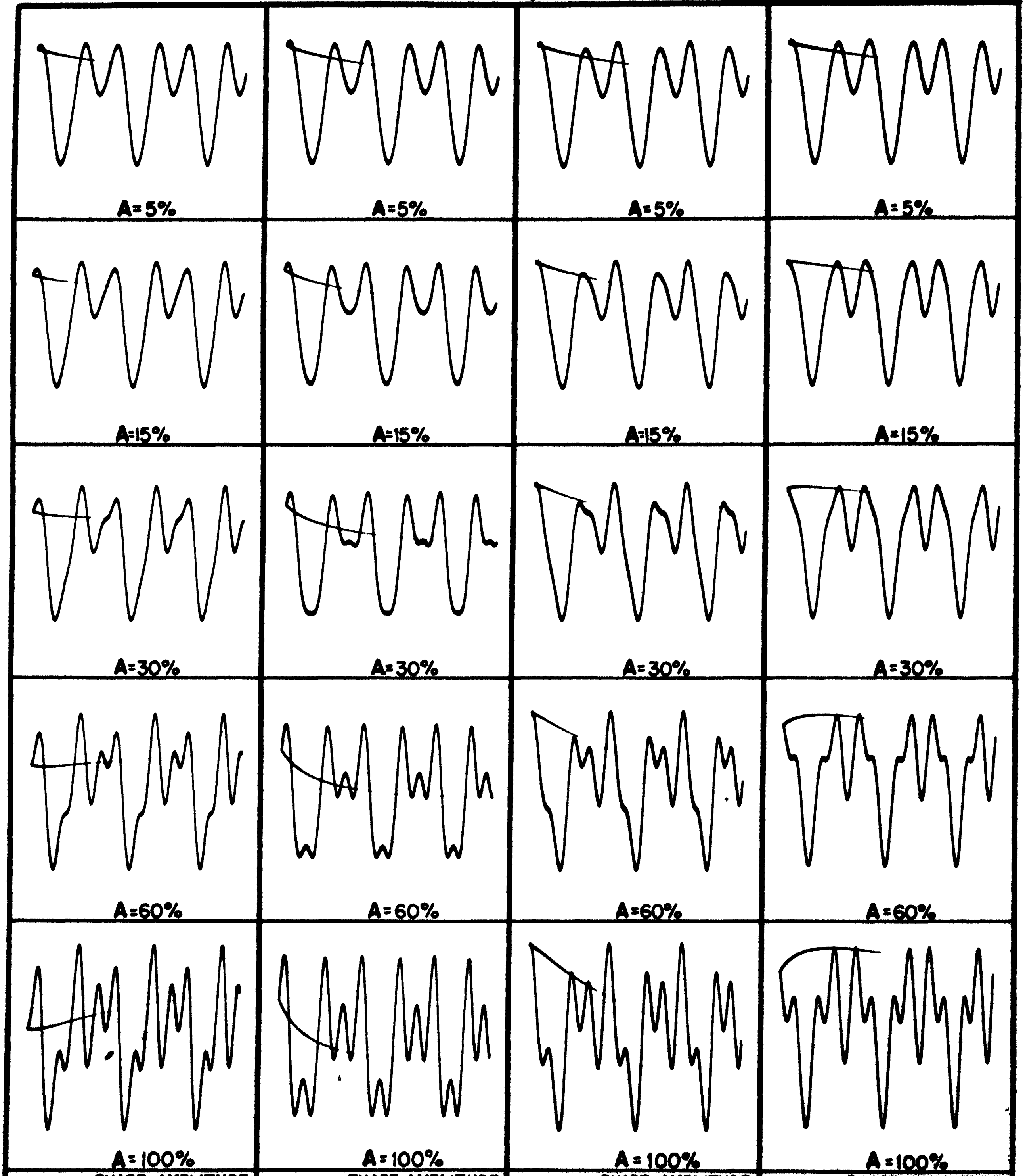
	PHASE	AMPLITUDE
Fund.	0°	100%
2nd	+90°	60%
4th	0°	*see* A

	PHASE	AMPLITUDE
Fund.	0°	100%
2nd	+90°	60%
4th	+90°	*see* A

	PHASE	AMPLITUDE
Fund.	0°	100%
2nd	+90°	60%
4th	180°	*see* A

	PHASE	AMPLITUDE
Fund.	0°	100%
2nd	+90°	60%
4th	-90°	*see* A

Fund. { 0° 100% + 2nd { +90° 100% + 4th { 0° to −90° 5% to 100%



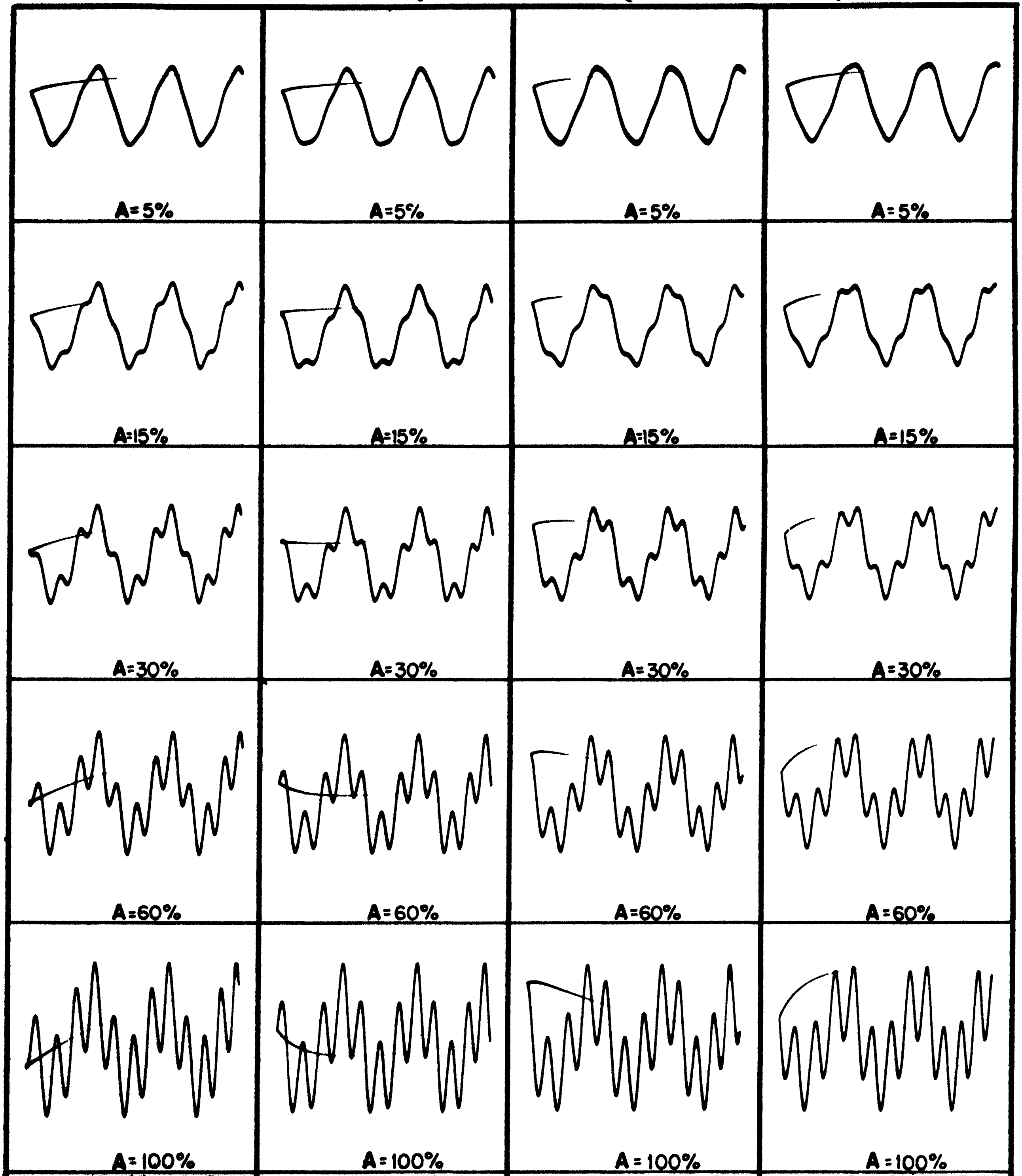
	PHASE	AMPLITUDE
Fund.	0°	100%
2nd	+90°	100%
4th	0°	see A

	PHASE	AMPLITUDE
Fund.	0°	100%
2nd	+90°	100%
4th	+90°	see A

	PHASE	AMPLITUDE
Fund.	0°	100%
2nd	+90°	100%
4th	180°	see A

	PHASE	AMPLITUDE
Fund.	0°	100%
2nd	+90°	100%
4th	−90°	see A

$$\text{Fund.}\begin{cases}0^\circ\\100\%\end{cases} + \text{2nd}\begin{cases}180^\circ\\5\%\end{cases} + \text{4th}\begin{cases}0^\circ \text{ to } -90^\circ\\5\% \text{ to } 100\%\end{cases}$$



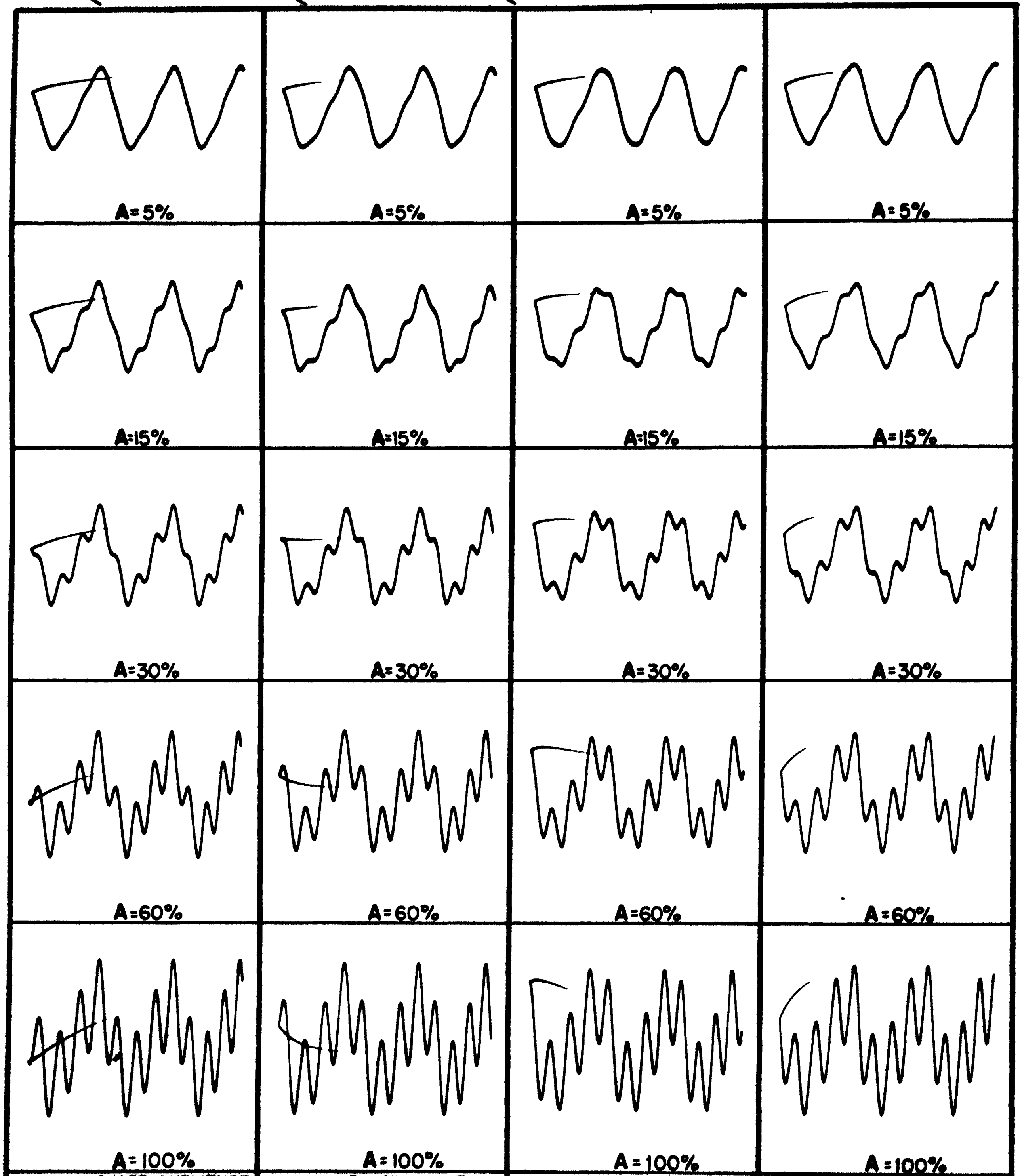
	PHASE	AMPLITUDE
Fund.	0°	100%
2nd	180°	5%
4th	0°	see A

	PHASE	AMPLITUDE
Fund.	0°	100%
2nd	180°	5%
4th	+90°	see A

	PHASE	AMPLITUDE
Fund.	0°	100%
2nd	180°	5%
4th	180°	see A

	PHASE	AMPLITUDE
Fund.	0°	100%
2nd	180°	5%
4th	-90°	see A

$$\text{Fund.}\left\{\begin{matrix}0^\circ\\100\%\end{matrix}\right. + \text{2nd}\left\{\begin{matrix}180^\circ\\15\%\end{matrix}\right. + \text{4th}\left\{\begin{matrix}0^\circ & \text{to} & -90^\circ\\5\% & \text{to} & 100\%\end{matrix}\right.$$



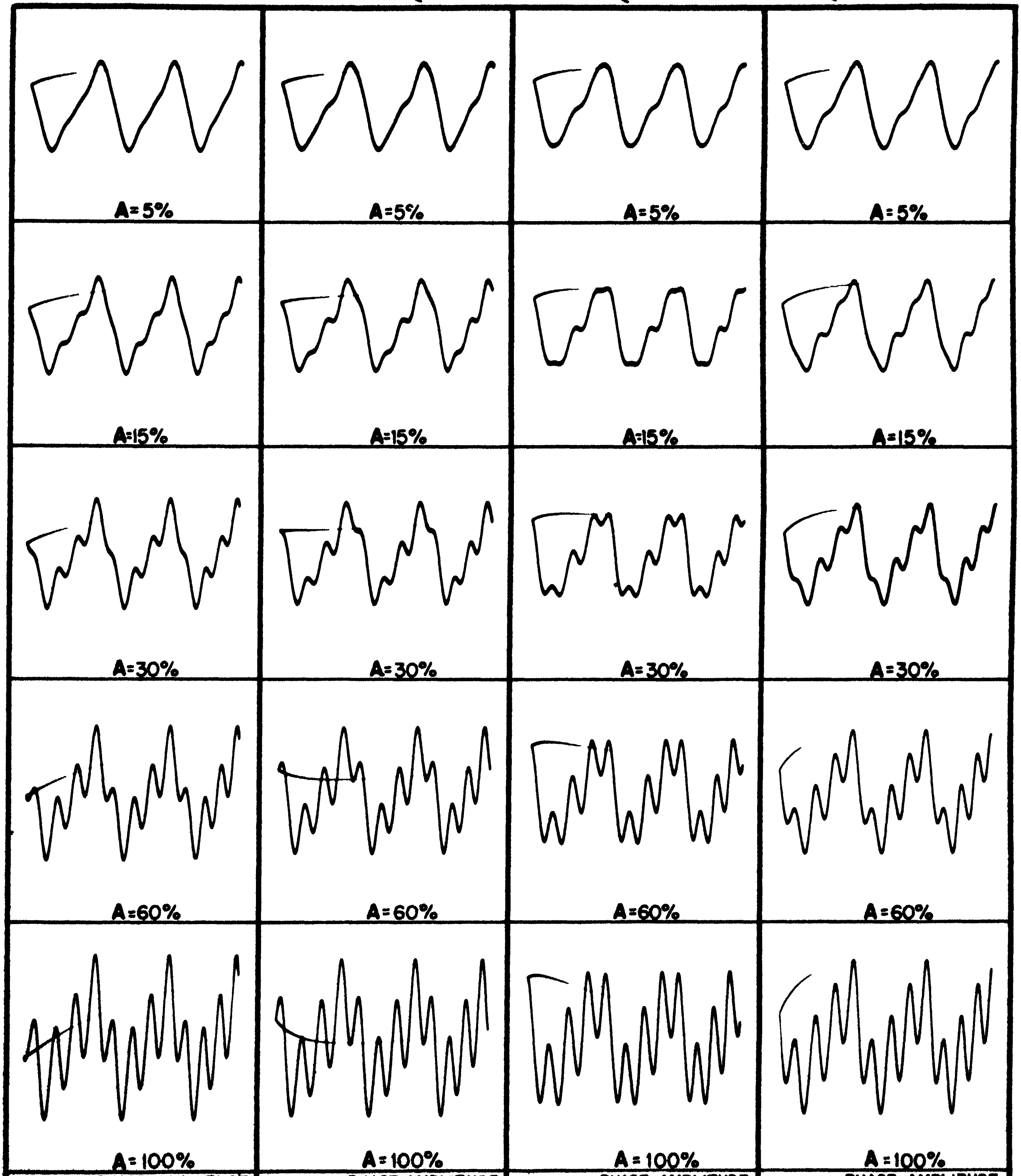
	PHASE	AMPLITUDE
Fund.	0°	100%
2nd	180°	15%
4th	0°	see A

	PHASE	AMPLITUDE
Fund.	0°	100%
2nd	180°	15%
4th	+90°	see A

	PHASE	AMPLITUDE
Fund.	0°	100%
2nd	180°	15%
4th	180°	see A

	PHASE	AMPLITUDE
Fund.	0°	100%
2nd	180°	15%
4th	-90°	see A

Fund. {0° 100% + **2nd** {180° 30% + **4th** {0° to −90° 5% to 100%



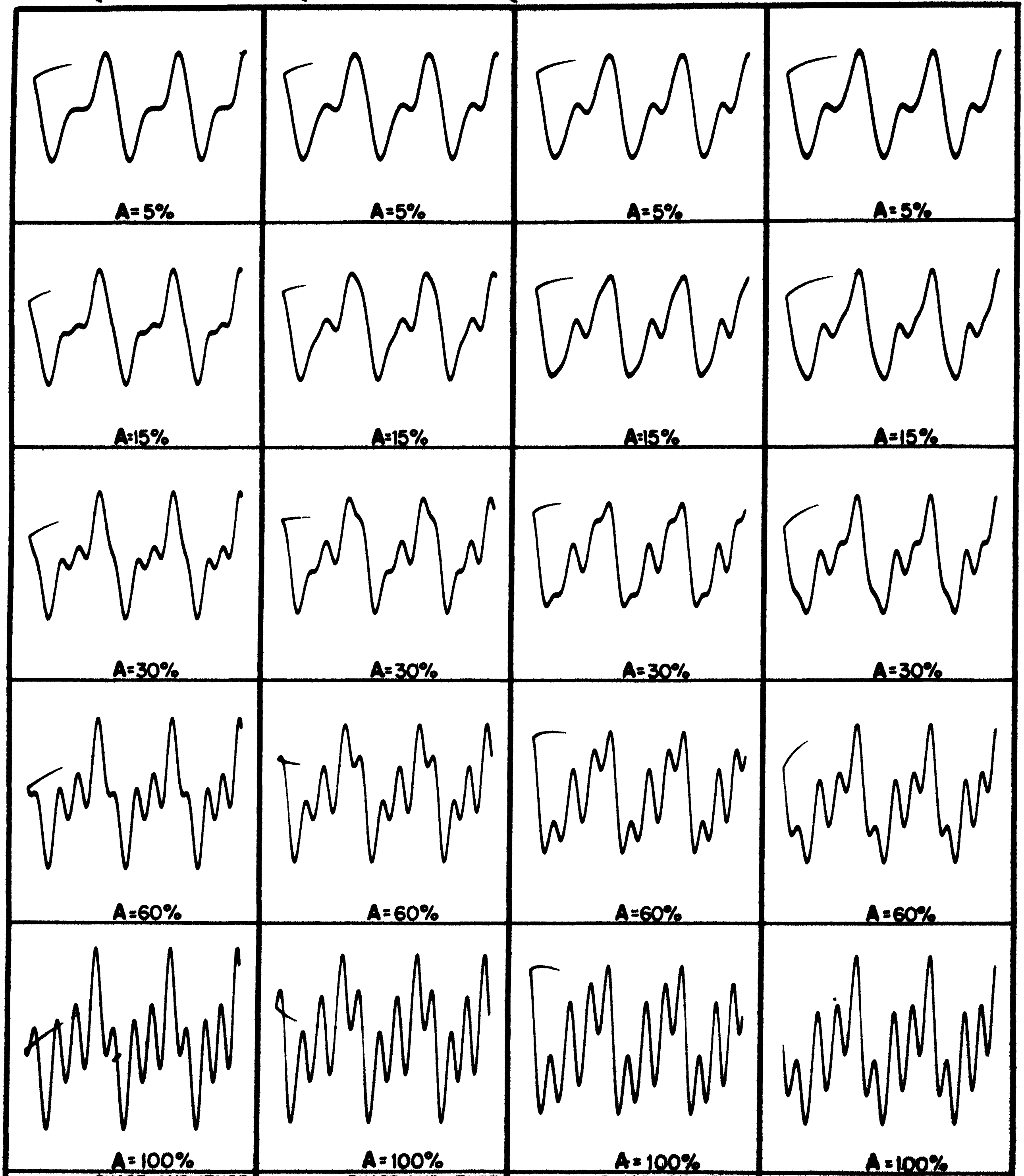
	PHASE	AMPLITUDE
Fund.	0°	100%
2nd	180°	30%
4th	0°	see A

	PHASE	AMPLITUDE
Fund.	0°	100%
2nd	180°	30%
4th	+90°	see A

	PHASE	AMPLITUDE
Fund.	0°	100%
2nd	180°	30%
4th	180°	see A

	PHASE	AMPLITUDE
Fund.	0°	100%
2nd	180°	30%
4th	−90°	see A

Fund. {0° 100% + **2nd** {180° 60% + **4th** {0° to −90° 5% to 100%



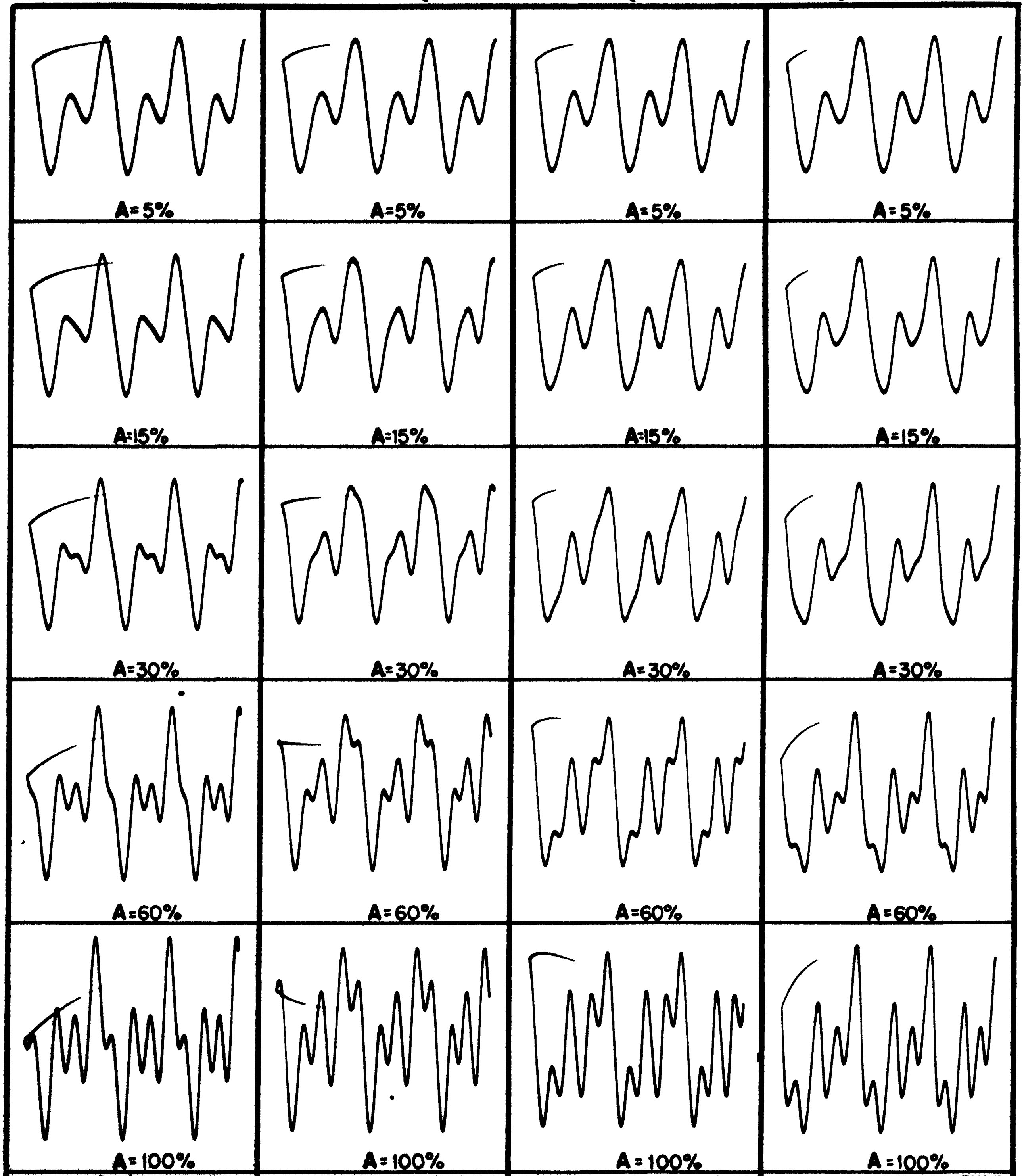
	PHASE	AMPLITUDE
Fund.	0°	100%
2nd	180°	60%
4th	0°	see A

	PHASE	AMPLITUDE
Fund.	0°	100%
2nd	180°	60%
4th	+90°	see A

	PHASE	AMPLITUDE
Fund.	0°	100%
2nd	180°	60%
4th	180°	see A

	PHASE	AMPLITUDE
Fund.	0°	100%
2nd	180°	60%
4th	-90°	see A

$$\text{Fund.}\left\{\begin{matrix}0^\circ\\100\%\end{matrix}\right. + \text{2nd}\left\{\begin{matrix}180^\circ\\100\%\end{matrix}\right. + \text{4th}\left\{\begin{matrix}0^\circ & \text{to} & -90^\circ\\5\% & \text{to} & 100\%\end{matrix}\right.$$



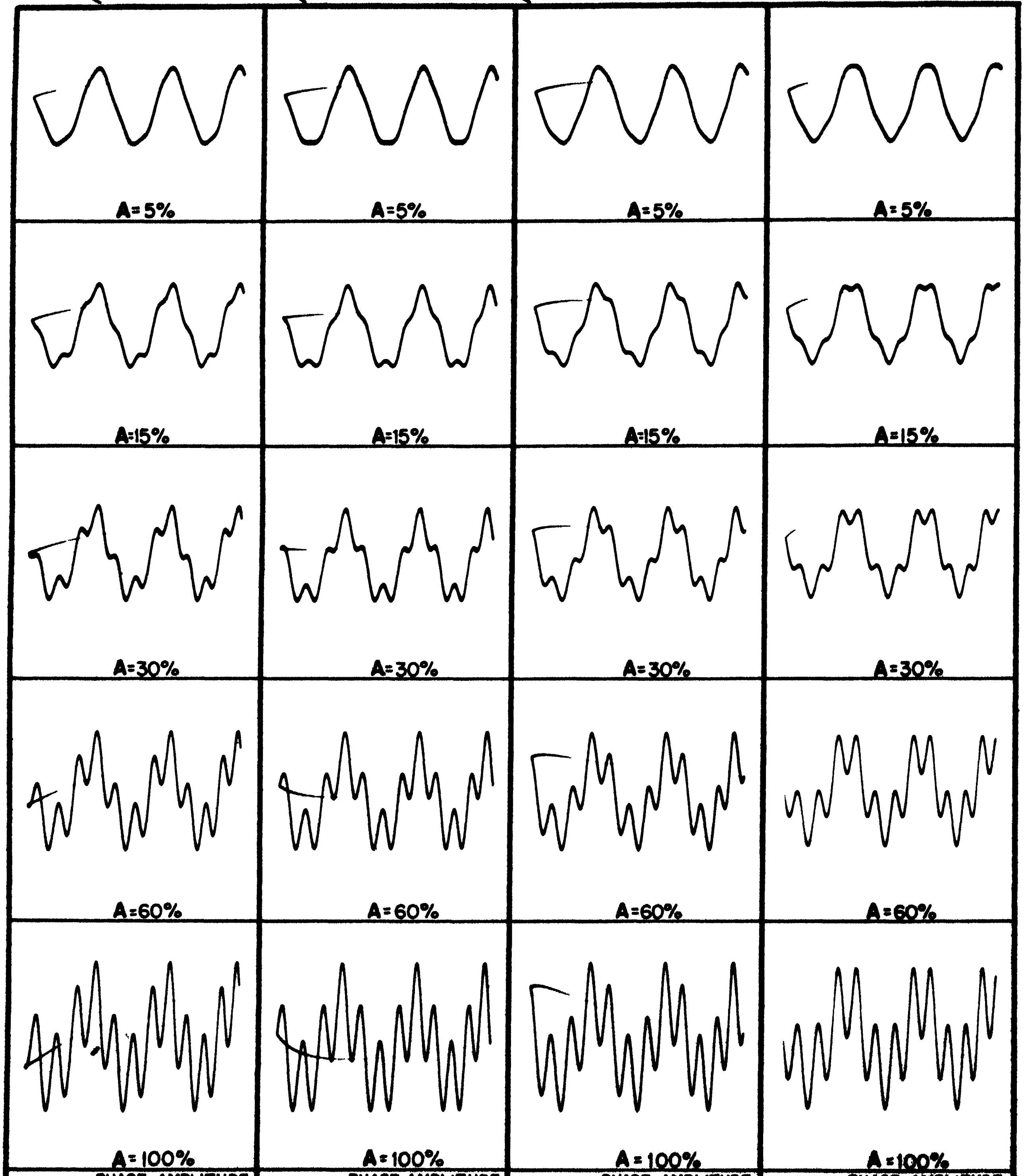
	PHASE	AMPLITUDE
Fund.	0°	100%
2nd	180°	100%
4th	0°	*see* A

	PHASE	AMPLITUDE
Fund.	0°	100%
2nd	180°	100%
4th	+90°	*see* A

	PHASE	AMPLITUDE
Fund.	0°	100%
2nd	180°	100%
4th	180°	*see* A

	PHASE	AMPLITUDE
Fund.	0°	100%
2nd	180°	100%
4th	-90°	*see* A

$$\text{Fund.}\left\{\begin{matrix}0^\circ\\ 100\%\end{matrix}\right. + \text{2nd}\left\{\begin{matrix}-90^\circ\\ 5\%\end{matrix}\right. + \text{4th}\left\{\begin{matrix}0^\circ \text{ to } -90^\circ\\ 5\% \text{ to } 100\%\end{matrix}\right.$$



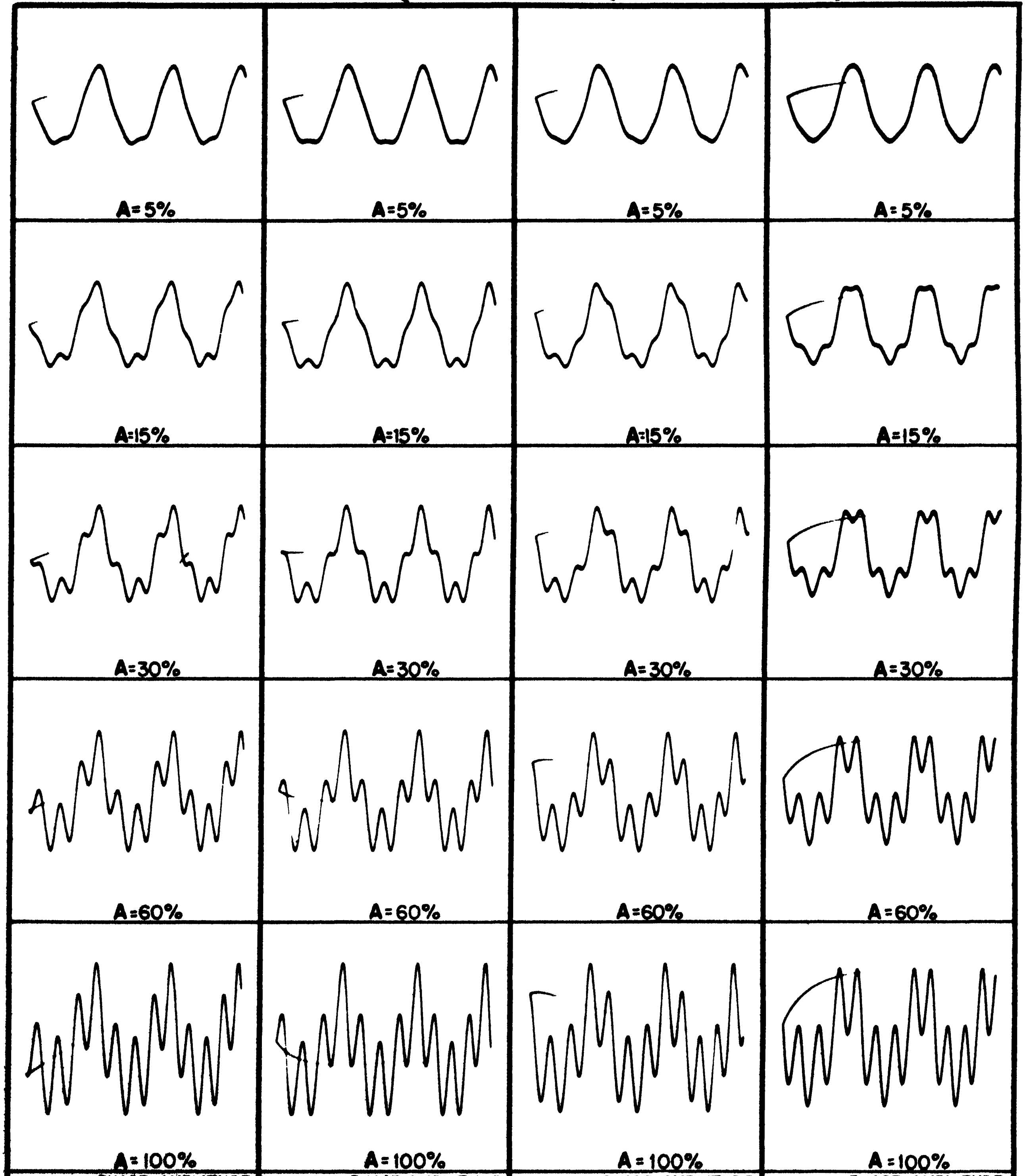
	PHASE	AMPLITUDE
Fund.	0°	100%
2nd	-90°	5%
4th	0°	*see* A

	PHASE	AMPLITUDE
Fund.	0°	100%
2nd	-90°	5%
4th	+90°	*see* A

	PHASE	AMPLITUDE
Fund.	0°	100%
2nd	-90°	5%
4th	180°	*see* A

	PHASE	AMPLITUDE
Fund.	0°	100%
2nd	-90°	5%
4th	-90°	*see* A

$$\text{Fund.}\left\{\begin{matrix}0^\circ\\100\%\end{matrix}\right. + \text{2nd}\left\{\begin{matrix}-90^\circ\\15\%\end{matrix}\right. + \text{4th}\left\{\begin{matrix}0^\circ & \text{to} & -90^\circ\\5\% & \text{to} & 100\%\end{matrix}\right.$$



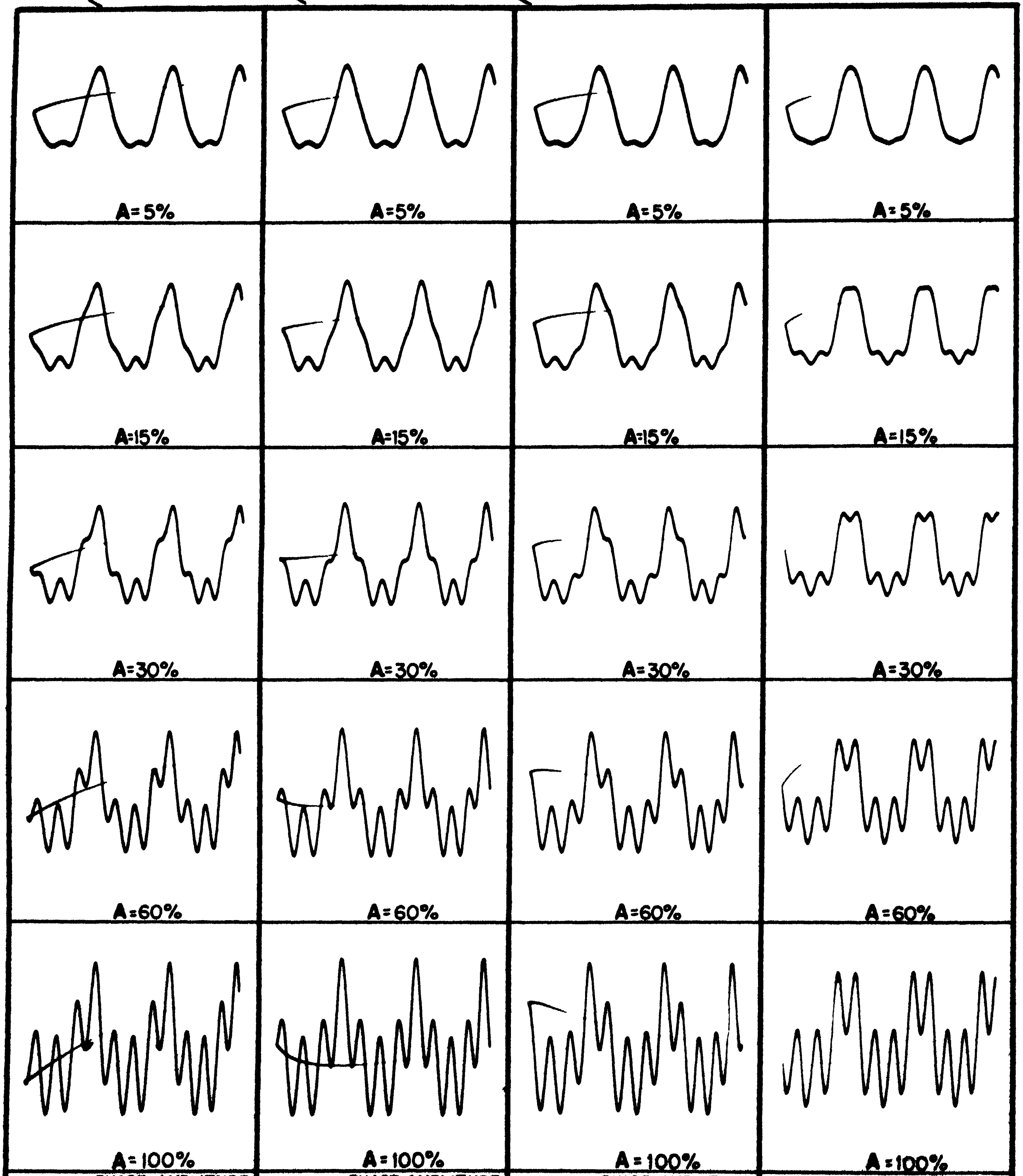
	PHASE	AMPLITUDE
Fund.	0°	100%
2nd	-90°	15%
4th	0°	see A

	PHASE	AMPLITUDE
Fund.	0°	100%
2nd	-90°	15%
4th	+90°	see A

	PHASE	AMPLITUDE
Fund.	0°	100%
2nd	-90°	15%
4th	180°	see A

	PHASE	AMPLITUDE
Fund.	0°	100%
2nd	-90°	15%
4th	-90°	see A

$$\text{Fund.}\begin{cases}0^\circ\\100\%\end{cases} + \text{2nd}\begin{cases}-90^\circ\\30\%\end{cases} + \text{4th}\begin{cases}0^\circ \text{ to } -90^\circ\\5\% \text{ to } 100\%\end{cases}$$



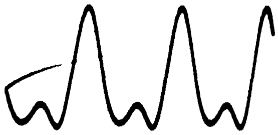
	PHASE	AMPLITUDE
Fund.	0°	100%
2nd	-90°	30%
4th	0°	see A

	PHASE	AMPLITUDE
Fund.	0°	100%
2nd	-90°	30%
4th	+90°	see A

	PHASE	AMPLITUDE
Fund.	0°	100%
2nd	-90°	30%
4th	180°	see A

	PHASE	AMPLITUDE
Fund.	0°	100%
2nd	-90°	30%
4th	-90°	see A

$$\text{Fund.}\begin{cases}0^\circ\\100\%\end{cases} + \text{2nd}\begin{cases}-90^\circ\\60\%\end{cases} + \text{4th}\begin{cases}0^\circ \text{ to } -90^\circ\\5\% \text{ to } 100\%\end{cases}$$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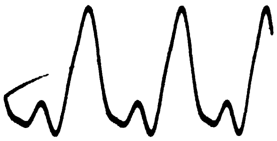
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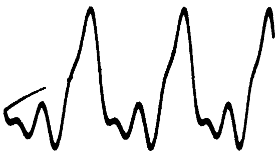
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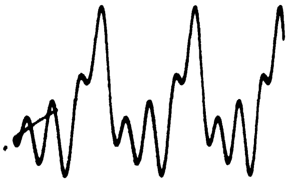
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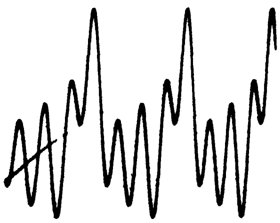
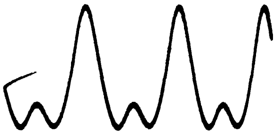

A=100%

	PHASE	AMPLITUDE
Fund.	0°	100%
2nd	-90°	60%
4th	0°	*see* A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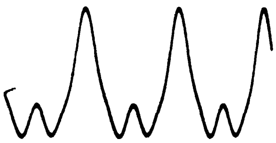
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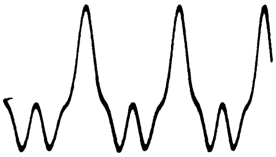
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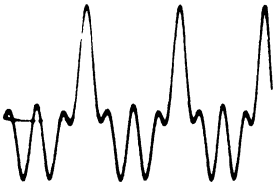
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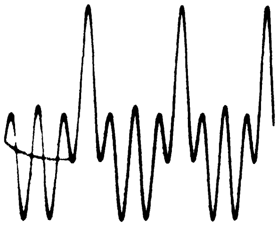
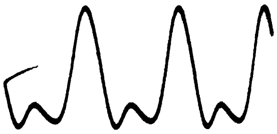

A=100%

	PHASE	AMPLITUDE
Fund.	0°	100%
2nd	-90°	60%
4th	+90°	*see* A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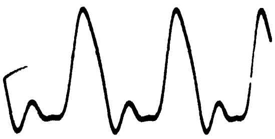
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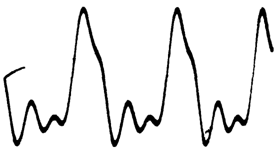
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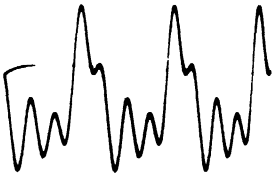
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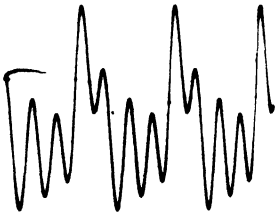
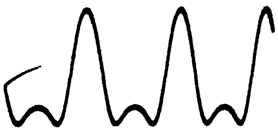

A=100%

	PHASE	AMPLITUDE
Fund.	0°	100%
2nd	-90°	60%
4th	180°	*see* A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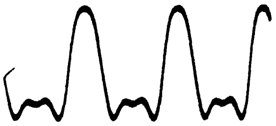
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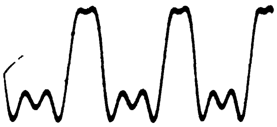
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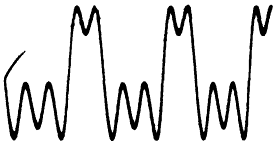
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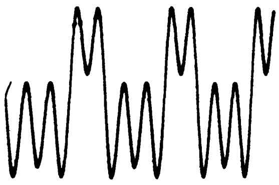
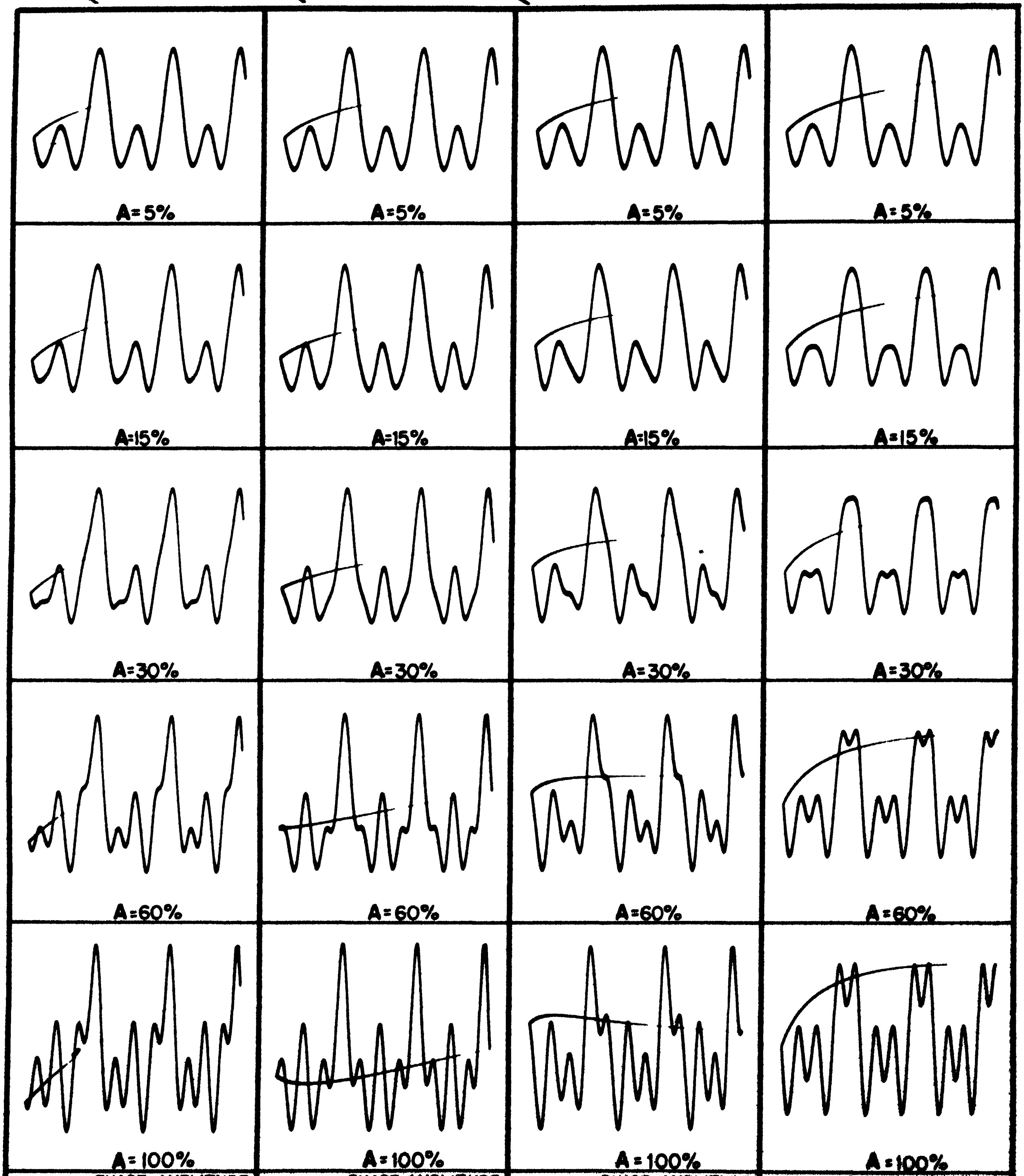

A=100%

	PHASE	AMPLITUDE
Fund.	0°	100%
2nd	-90°	60%
4th	-90°	*see* A

Fund. $\left\{\begin{matrix} 0° \\ 100\% \end{matrix}\right.$ + **2nd** $\left\{\begin{matrix} -90° \\ 100\% \end{matrix}\right.$ + **4th** $\left\{\begin{matrix} 0° \text{ to } -90° \\ 5\% \text{ to } 100\% \end{matrix}\right.$



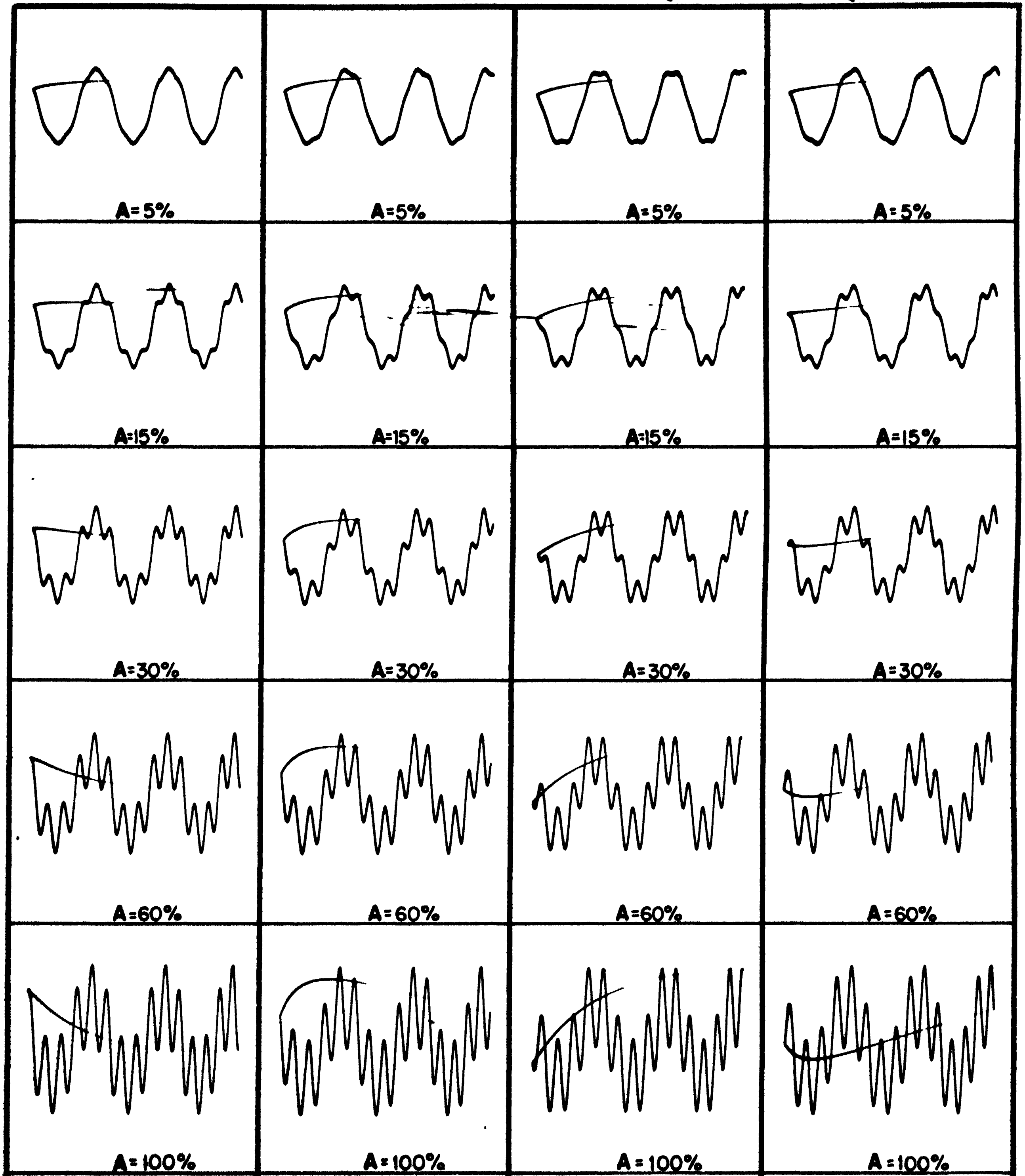
	PHASE	AMPLITUDE
Fund.	0°	100%
2nd	-90°	100%
4th	0°	*see* A

	PHASE	AMPLITUDE
Fund.	0°	100%
2nd	-90°	100%
4th	+90°	*see* A

	PHASE	AMPLITUDE
Fund.	0°	100%
2nd	-90°	100%
4th	180°	*see* A

	PHASE	AMPLITUDE
Fund.	0°	100%
2nd	-90°	100%
4th	-90°	*see* A

$$\text{Fund.}\begin{cases}0^\circ\\100\%\end{cases} + \text{3rd}\begin{cases}0^\circ\\5\%\end{cases} + \text{5th}\begin{cases}0^\circ \text{ to } -90^\circ\\5\% \text{ to } 100\%\end{cases}$$



A=5% A=5% A=5% A=5%

A=15% A=15% A=15% A=15%

A=30% A=30% A=30% A=30%

A=60% A=60% A=60% A=60%

A=100% A=100% A=100% A=100%

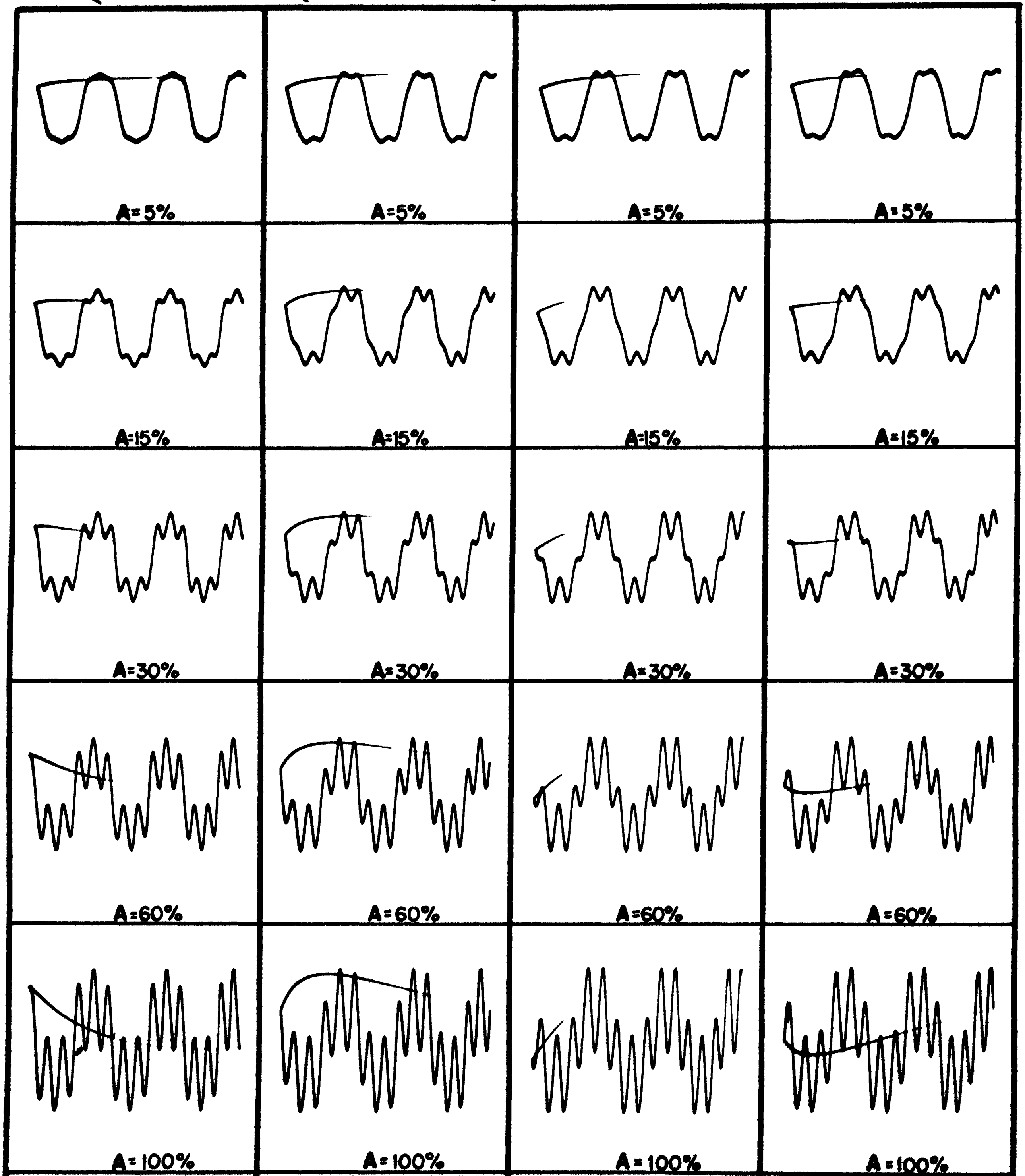
	PHASE	AMPLITUDE
Fund.	0°	100%
3rd	0°	5%
5th	0°	see A

	PHASE	AMPLITUDE
Fund.	0°	100%
3rd	0°	5%
5th	+90°	see A

	PHASE	AMPLITUDE
Fund.	0°	100%
3rd	0°	5%
5th	180°	see A

	PHASE	AMPLITUDE
Fund.	0°	100%
3rd	0°	5%
5th	-90°	see A

Fund. {0° 100% + 3rd {0° 15% + 5th {0° to −90° 5% to 100%



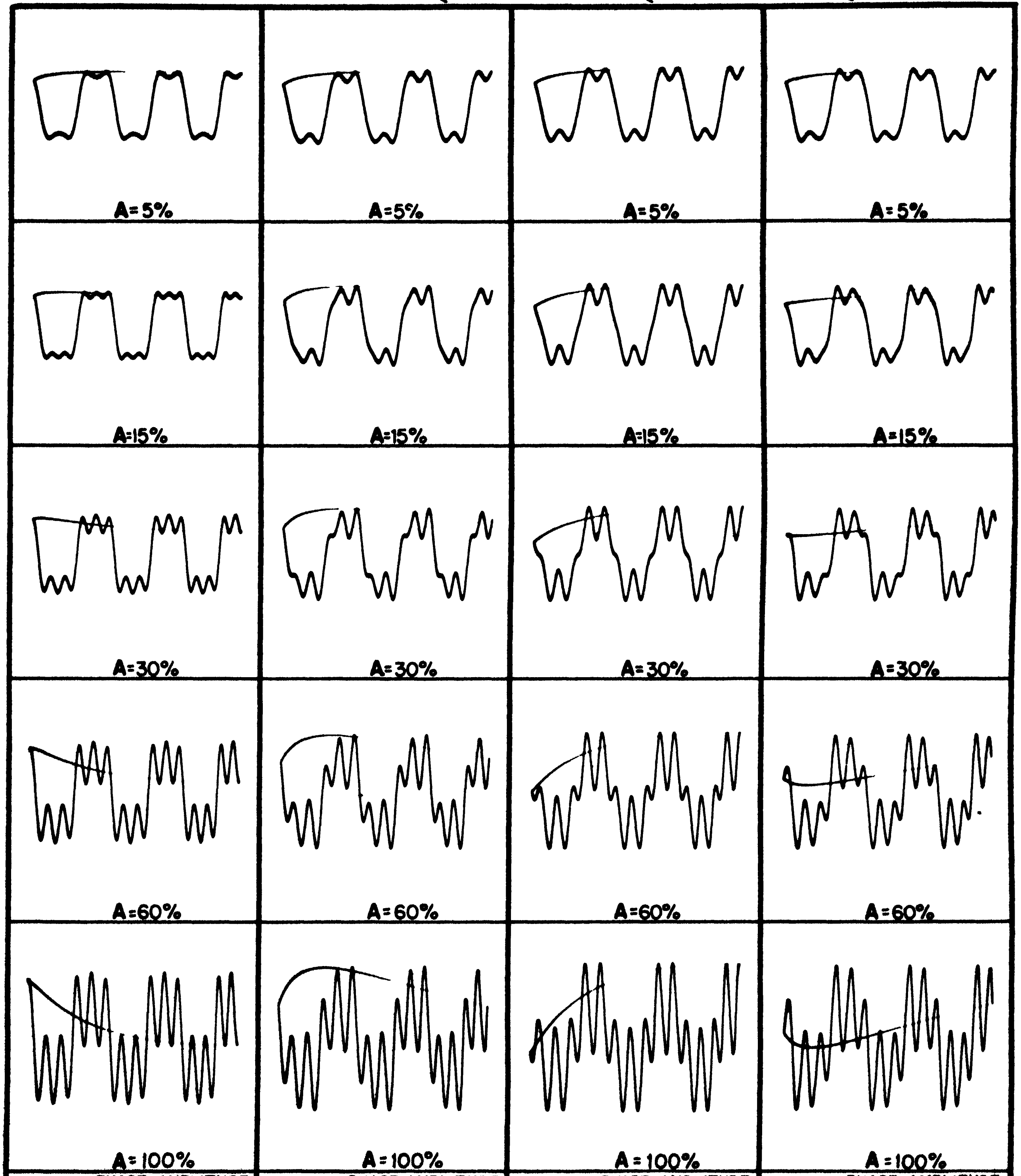
	PHASE	AMPLITUDE
Fund.	0°	100%
3rd	0°	15%
5th	0°	see A

	PHASE	AMPLITUDE
Fund.	0°	100%
3rd	0°	15%
5th	+90°	see A

	PHASE	AMPLITUDE
Fund.	0°	100%
3rd	0°	15%
5th	180°	see A

	PHASE	AMPLITUDE
Fund.	0°	100%
3rd	0°	15%
5th	−90°	see A

Fund. $\left\{\begin{matrix}0°\\100\%\end{matrix}\right.$ + 3rd $\left\{\begin{matrix}0°\\30\%\end{matrix}\right.$ + 5th $\left\{\begin{matrix}0° \text{ to } -90°\\5\% \text{ to } 100\%\end{matrix}\right.$



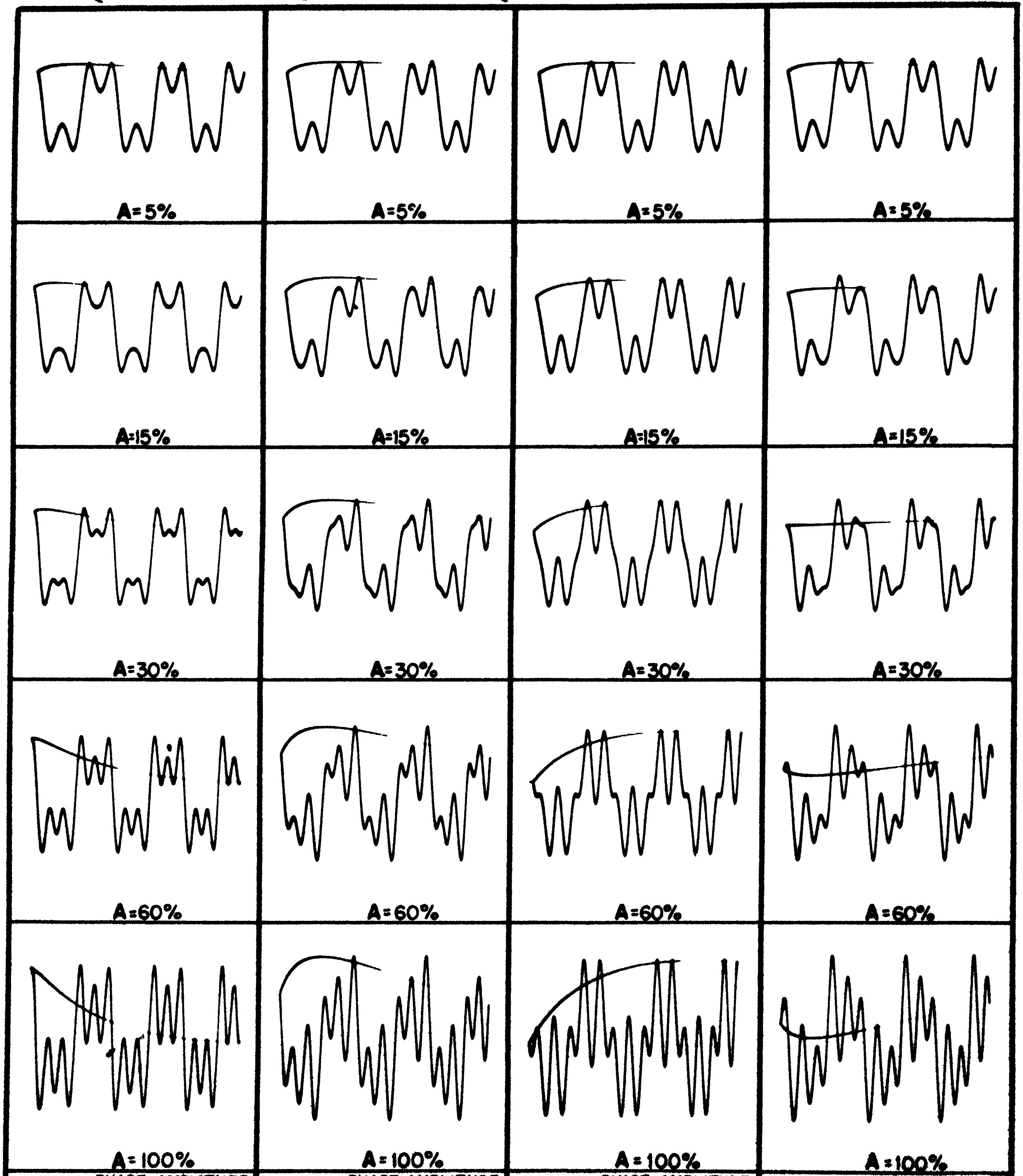
	PHASE	AMPLITUDE
Fund.	0°	100%
3rd	0°	30%
5th	0°	*see* A

	PHASE	AMPLITUDE
Fund.	0°	100%
3rd	0°	30%
5th	+90°	*see* A

	PHASE	AMPLITUDE
Fund.	0°	100%
3rd	0°	30%
5th	180°	*see* A

	PHASE	AMPLITUDE
Fund.	0°	100%
3rd	0°	30%
5th	-90°	*see* A

$$\text{Fund.}\begin{cases}0^\circ\\100\%\end{cases} + \text{3rd}\begin{cases}0^\circ\\60\%\end{cases} + \text{5th}\begin{cases}0^\circ \text{ to } -90^\circ\\5\% \text{ to } 100\%\end{cases}$$



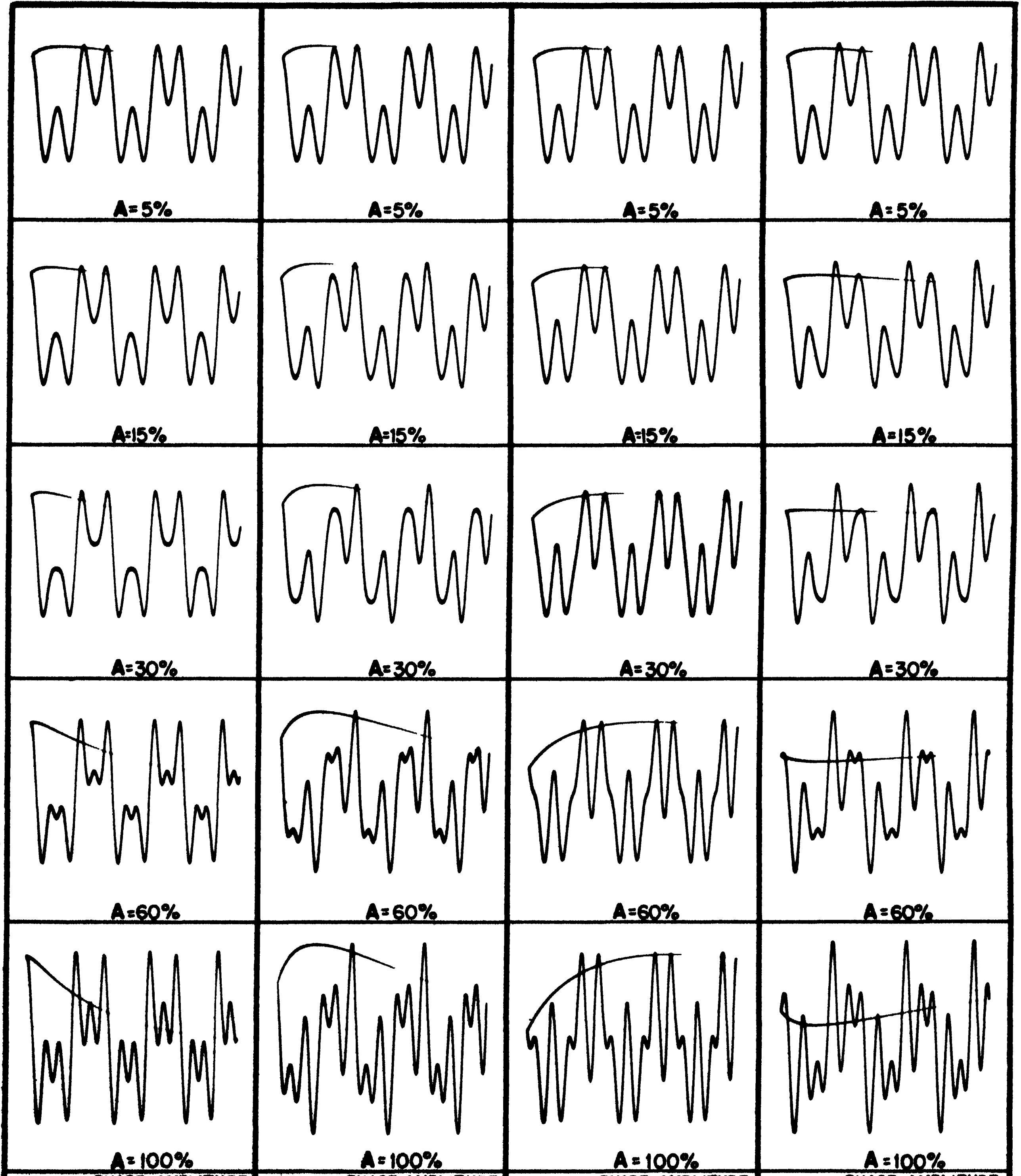
	PHASE	AMPLITUDE
Fund.	0°	100%
3rd	0°	60%
5th	0°	see A

	PHASE	AMPLITUDE
Fund.	0°	100%
3rd	0°	60%
5th	+90°	see A

	PHASE	AMPLITUDE
Fund.	0°	100%
3rd	0°	60%
5th	180°	see A

	PHASE	AMPLITUDE
Fund.	0°	100%
3rd	0°	60%
5th	-90°	see A

Fund. $\begin{cases} 0° \\ 100\% \end{cases}$ + 3rd $\begin{cases} 0° \\ 100\% \end{cases}$ + 5th $\begin{cases} 0° \text{ to } -90° \\ 5\% \text{ to } 100\% \end{cases}$



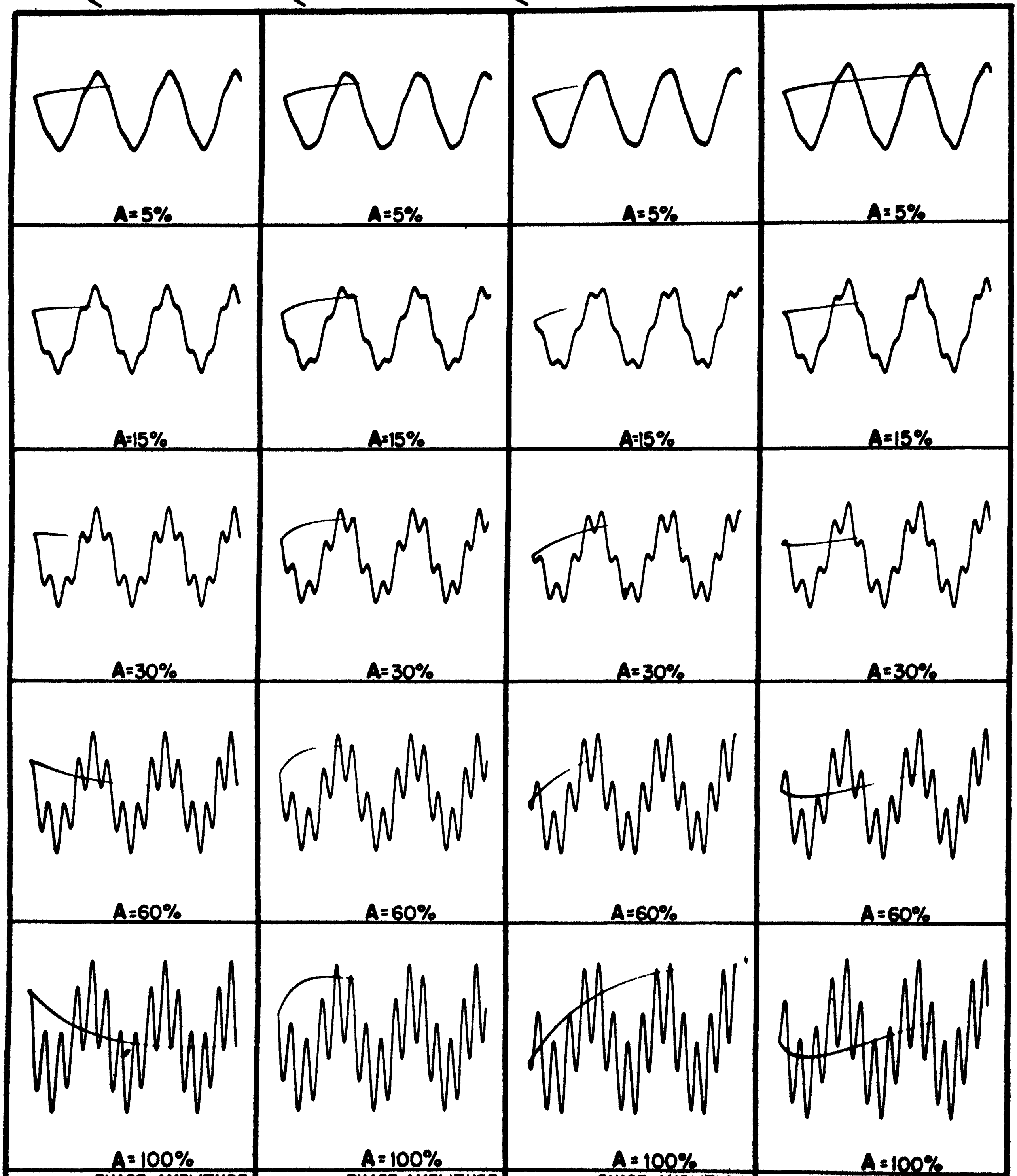
	PHASE	AMPLITUDE
Fund.	0°	100%
3rd	0°	100%
5th	0°	see A

	PHASE	AMPLITUDE
Fund.	0°	100%
3rd	0°	100%
5th	+90°	see A

	PHASE	AMPLITUDE
Fund.	0°	100%
3rd	0°	100%
5th	180°	see A

	PHASE	AMPLITUDE
Fund.	0°	100%
3rd	0°	100%
5th	-90°	see A

$$\text{Fund.}\begin{cases}0^\circ\\100\%\end{cases}+\ \text{3rd}\begin{cases}+90^\circ\\5\%\end{cases}+\ \text{5th}\begin{cases}0^\circ\text{ to }-90^\circ\\5\%\text{ to }100\%\end{cases}$$



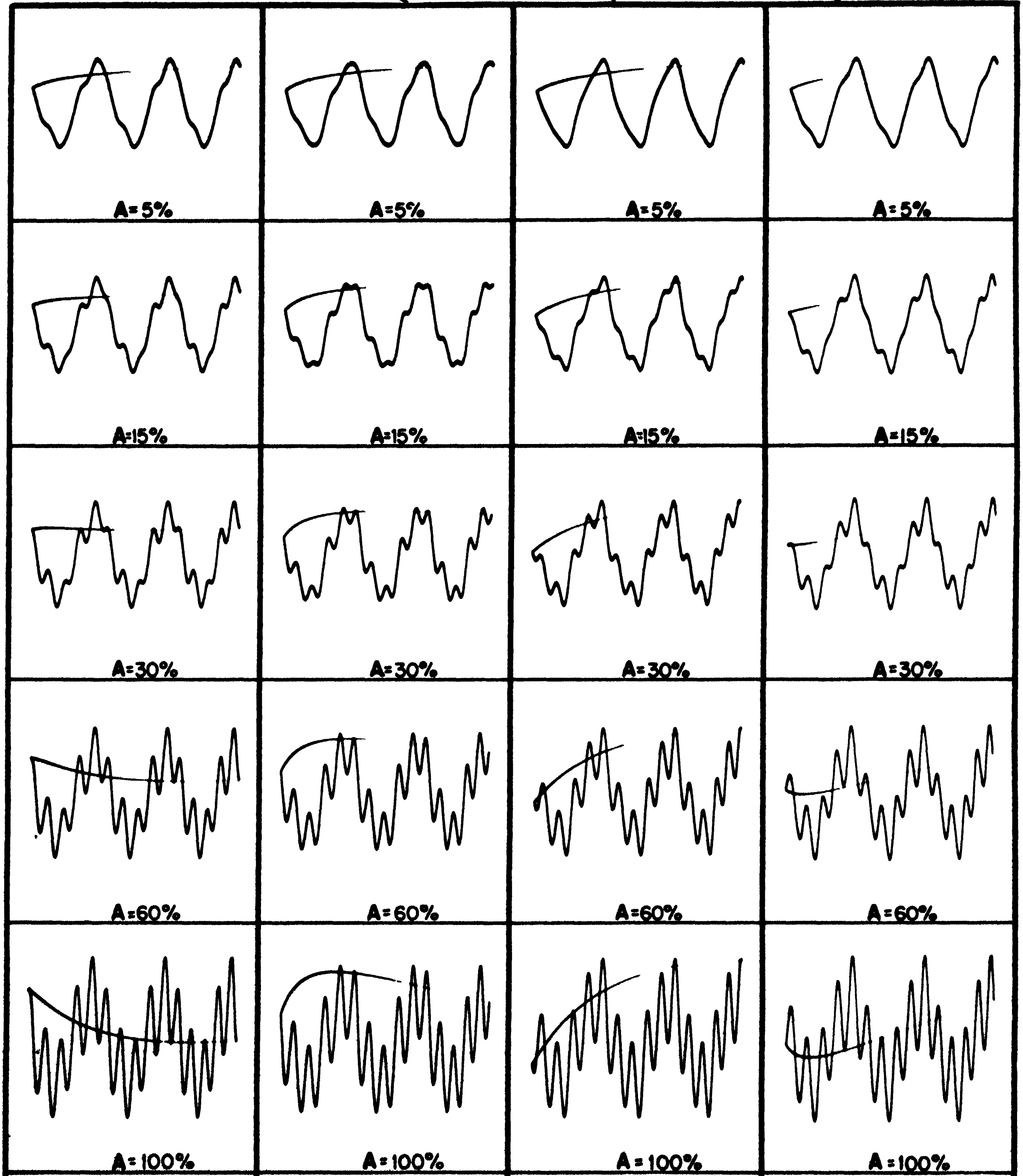
	PHASE	AMPLITUDE
Fund.	0°	100%
3rd	+90°	5%
5th	0°	see A

	PHASE	AMPLITUDE
Fund.	0°	100%
3rd	+90°	5%
5th	+90°	see A

	PHASE	AMPLITUDE
Fund.	0°	100%
3rd	+90°	5%
5th	180°	see A

	PHASE	AMPLITUDE
Fund.	0°	100%
3rd	+90°	5%
5th	-90°	see A

Fund. { 0° 100% + 3rd { +90° 15% + 5th { 0° to -90° 5% to 100%



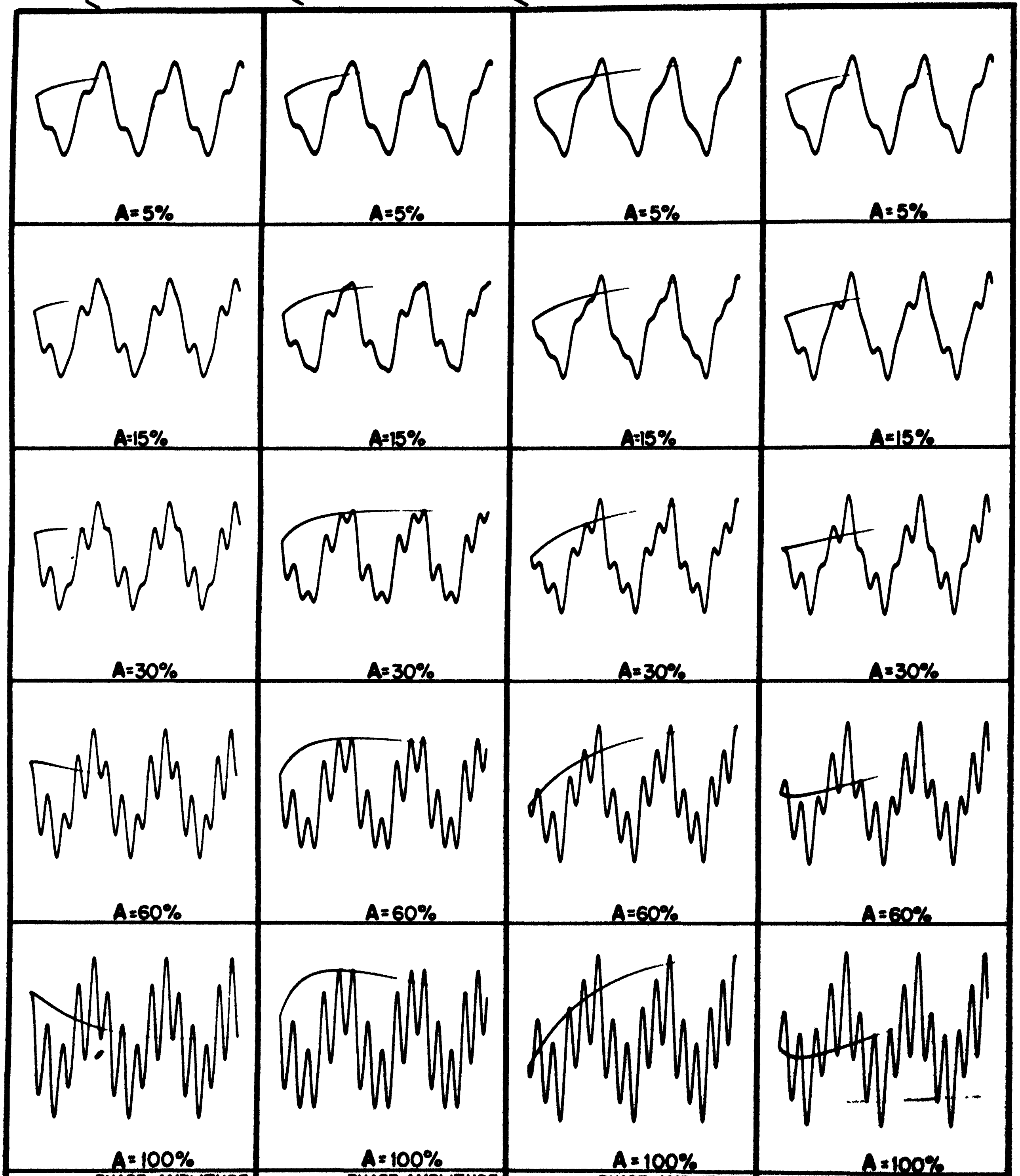
	PHASE	AMPLITUDE
Fund.	0°	100%
3rd	+90°	15%
5th	0°	see A

	PHASE	AMPLITUDE
Fund.	0°	100%
3rd	+90°	15%
5th	+90°	see A

	PHASE	AMPLITUDE
Fund.	0°	100%
3rd	+90°	15%
5th	180°	see A

	PHASE	AMPLITUDE
Fund.	0°	100%
3rd	+90°	15%
5th	-90°	see A

$$\text{Fund.}\begin{cases}0^\circ\\100\%\end{cases} + \text{3rd}\begin{cases}+90^\circ\\30\%\end{cases} + \text{5th}\begin{cases}0^\circ \text{ to } -90^\circ\\5\% \text{ to } 100\%\end{cases}$$



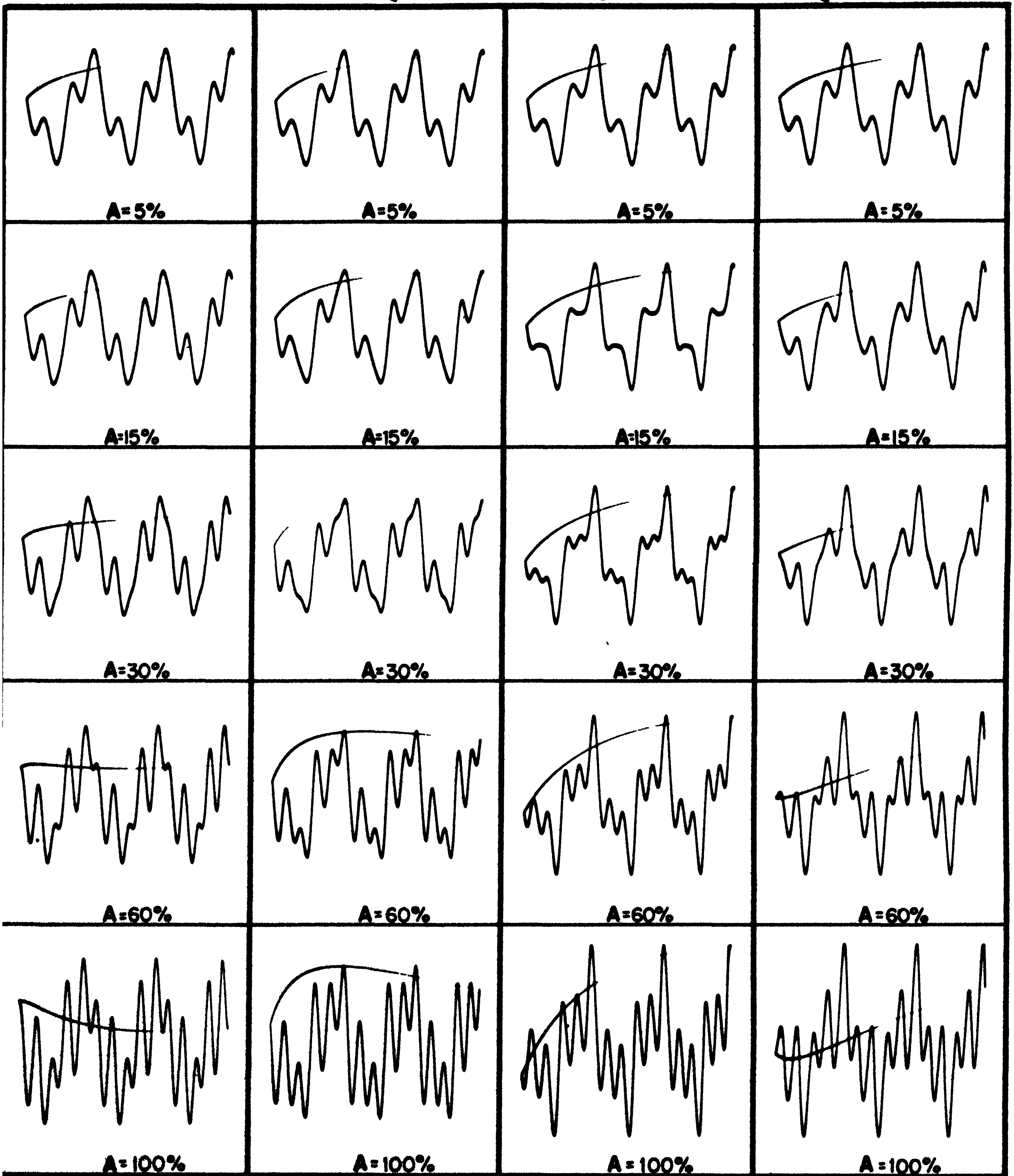
	PHASE	AMPLITUDE
Fund.	0°	100%
3rd	+90°	30%
5th	0°	see A

	PHASE	AMPLITUDE
Fund.	0°	100%
3rd	+90°	30%
5th	+90°	see A

	PHASE	AMPLITUDE
Fund.	0°	100%
3rd	+90°	30%
5th	180°	see A

	PHASE	AMPLITUDE
Fund.	0°	100%
3rd	+90°	30%
5th	-90°	see A

$$\text{Fund.}\begin{cases}0^\circ\\100\%\end{cases} + \text{3rd}\begin{cases}+90^\circ\\60\%\end{cases} + \text{5th}\begin{cases}0^\circ \text{ to } -90^\circ\\5\% \text{ to } 100\%\end{cases}$$



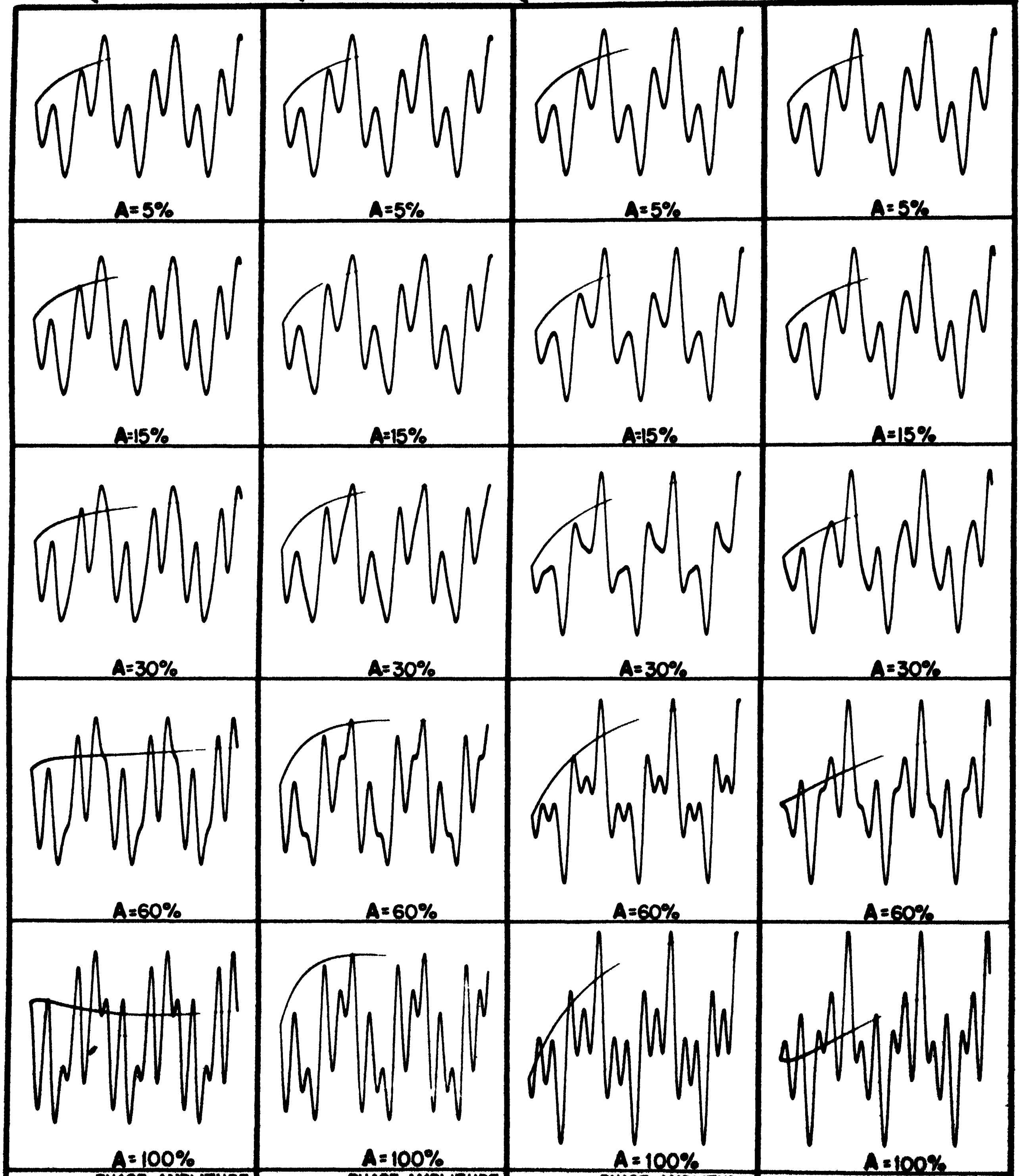
	PHASE	AMPLITUDE
Fund.	0°	100%
3rd	+90°	60%
5th	0°	*see* A

	PHASE	AMPLITUDE
Fund.	0°	100%
3rd	+90°	60%
5th	+90°	*see* A

	PHASE	AMPLITUDE
Fund.	0°	100%
3rd	+90°	60%
5th	180°	*see* A

	PHASE	AMPLITUDE
Fund.	0°	100%
3rd	+90°	60%
5th	-90°	*see* A

Fund. { 0° 100% + 3rd { +90° 100% + 5th { 0° to −90° 5% to 100%



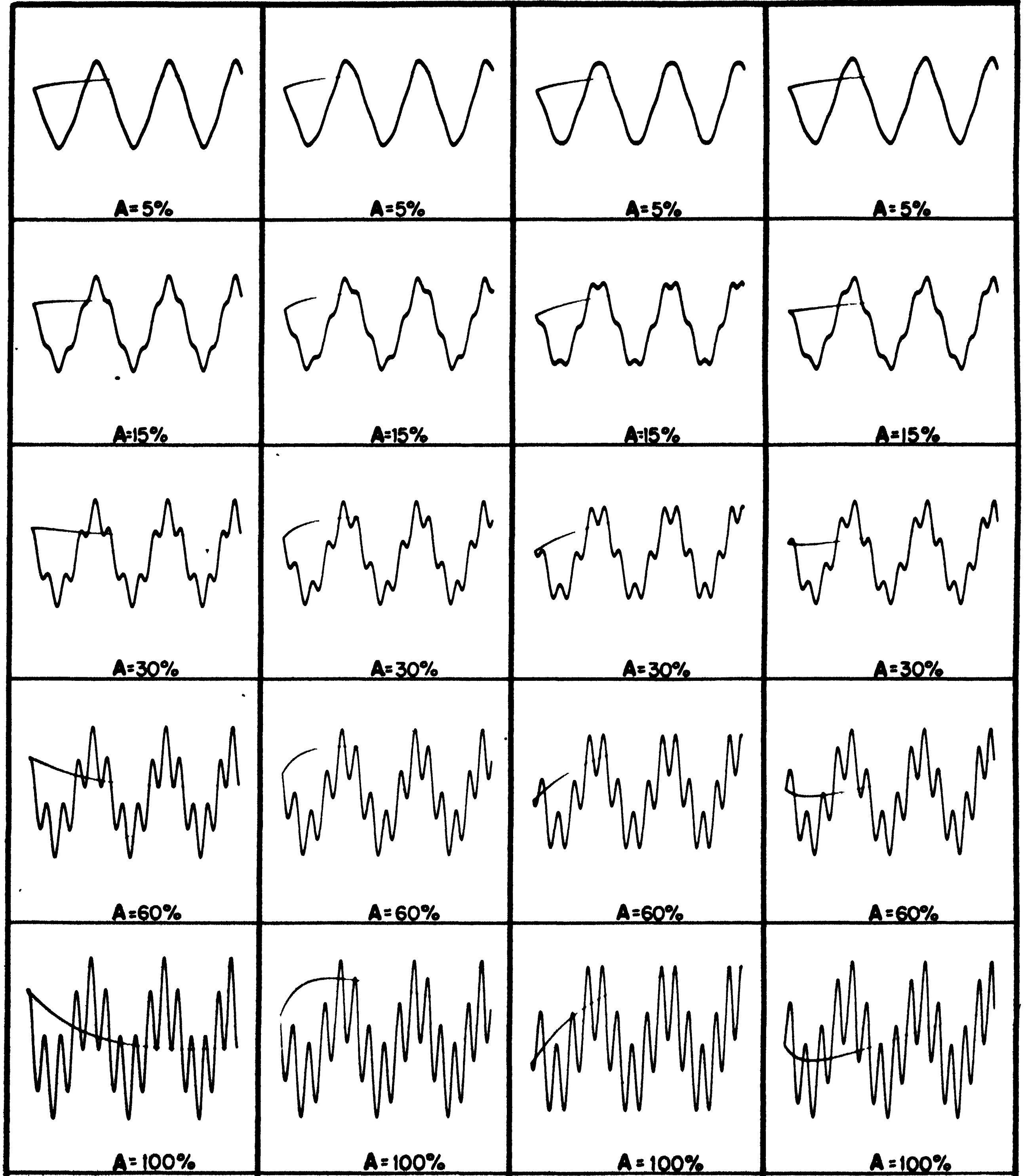
	PHASE	AMPLITUDE
Fund.	0°	100%
3rd	+90°	100%
5th	0°	*see A*

	PHASE	AMPLITUDE
Fund.	0°	100%
3rd	+90°	100%
5th	+90°	*see A*

	PHASE	AMPLITUDE
Fund.	0°	100%
3rd	+90°	100%
5th	180°	*see A*

	PHASE	AMPLITUDE
Fund.	0°	100%
3rd	+90°	100%
5th	-90°	*see A*

$$\text{Fund.}\left\{\begin{matrix}0^\circ\\100\%\end{matrix}\right. + \text{3rd}\left\{\begin{matrix}180^\circ\\5\%\end{matrix}\right. + \text{5th}\left\{\begin{matrix}0^\circ & \text{to} & -90^\circ\\5\% & \text{to} & 100\%\end{matrix}\right.$$



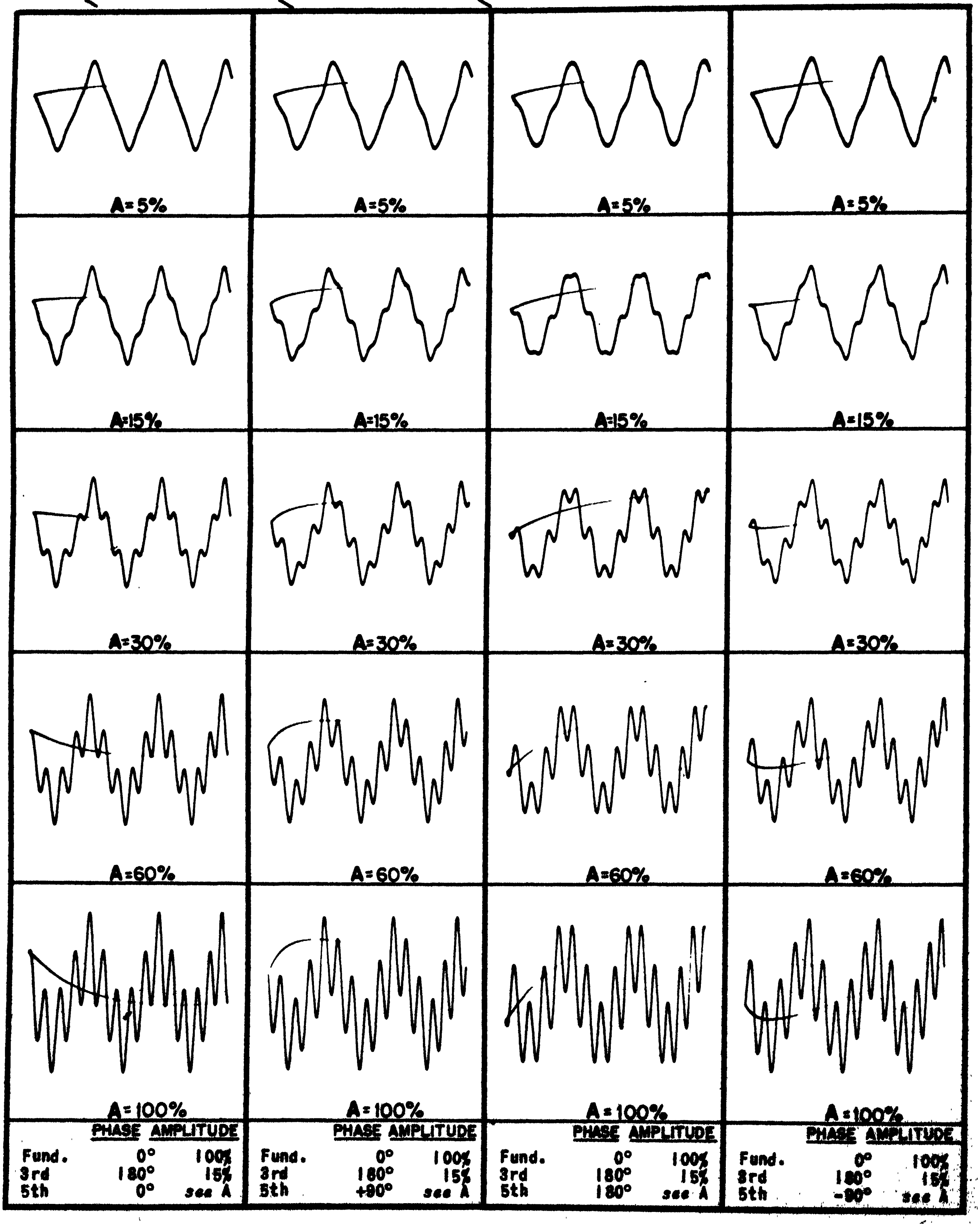
	PHASE	AMPLITUDE
Fund.	0°	100%
3rd	180°	5%
5th	0°	*see* A

	PHASE	AMPLITUDE
Fund.	0°	100%
3rd	180°	5%
5th	+90°	*see* A

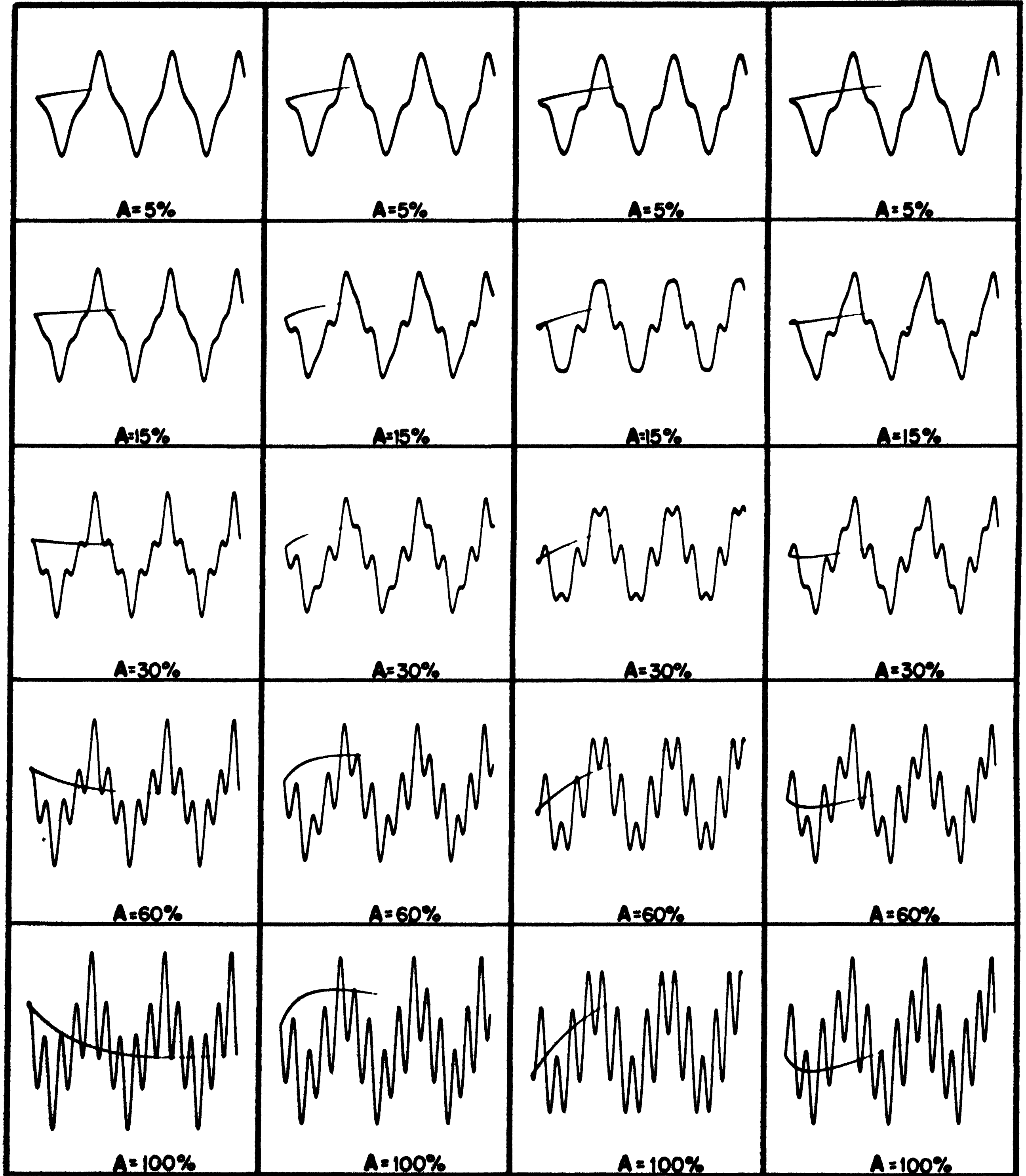
	PHASE	AMPLITUDE
Fund.	0°	100%
3rd	180°	5%
5th	180°	*see* A

	PHASE	AMPLITUDE
Fund.	0°	100%
3rd	180°	5%
5th	-90°	*see* A

Fund. $\begin{cases} 0° \\ 100\% \end{cases}$ + 3rd $\begin{cases} 180° \\ 15\% \end{cases}$ + 5th $\begin{cases} 0° \text{ to } -90° \\ 5\% \text{ to } 100\% \end{cases}$



$$\text{Fund.}\left\{\begin{array}{l}0^\circ\\100\%\end{array}\right. + \text{3rd}\left\{\begin{array}{l}180^\circ\\30\%\end{array}\right. + \text{5th}\left\{\begin{array}{lll}0^\circ & \text{to} & -90^\circ\\5\% & \text{to} & 100\%\end{array}\right.$$



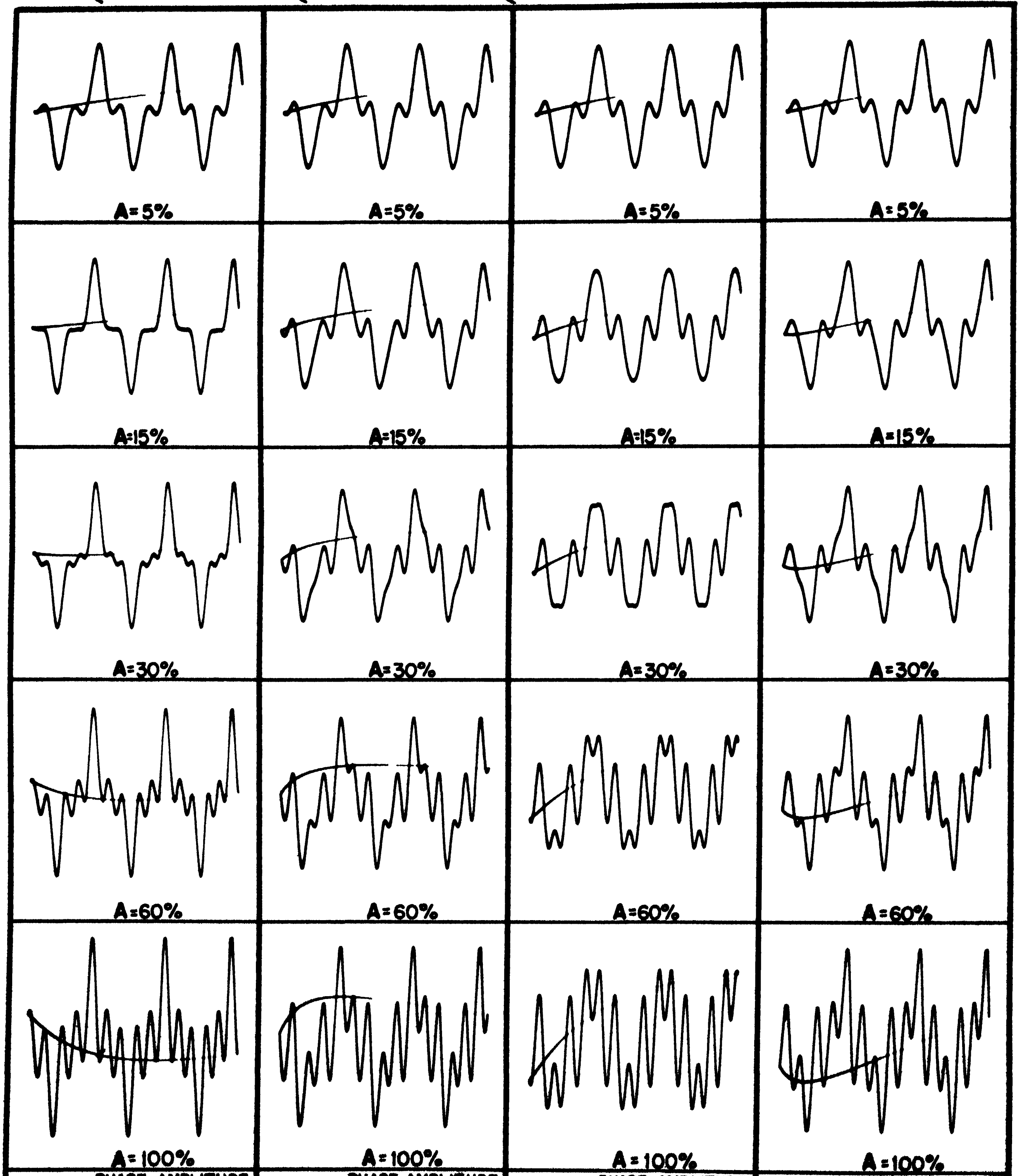
	PHASE	AMPLITUDE
Fund.	0°	100%
3rd	180°	30%
5th	0°	see A

	PHASE	AMPLITUDE
Fund.	0°	100%
3rd	180°	30%
5th	+90°	see A

	PHASE	AMPLITUDE
Fund.	0°	100%
3rd	180°	30%
5th	180°	see A

	PHASE	AMPLITUDE
Fund.	0°	100%
3rd	180°	30%
5th	-90°	see A

Fund. $\left\{\begin{matrix}0^\circ\\100\%\end{matrix}\right.$ + **3rd** $\left\{\begin{matrix}180^\circ\\60\%\end{matrix}\right.$ + **5th** $\left\{\begin{matrix}0^\circ \text{ to } -90^\circ\\5\% \text{ to } 100\%\end{matrix}\right.$



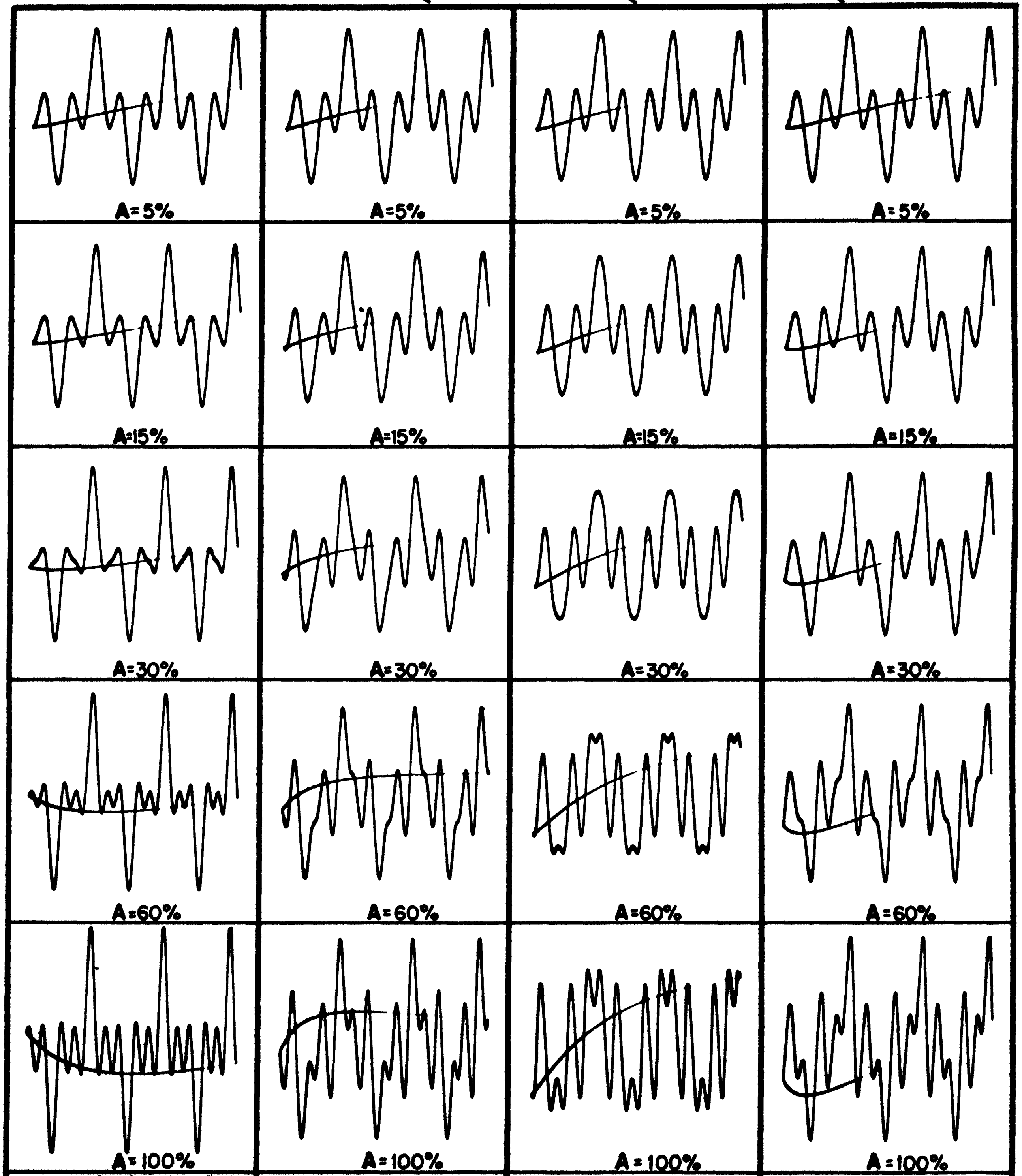
	PHASE	AMPLITUDE
Fund.	0°	100%
3rd	180°	60%
5th	0°	see A

	PHASE	AMPLITUDE
Fund.	0°	100%
3rd	180°	60%
5th	+90°	see A

	PHASE	AMPLITUDE
Fund.	0°	100%
3rd	180°	60%
5th	180°	see A

	PHASE	AMPLITUDE
Fund.	0°	100%
3rd	180°	60%
5th	-90°	see A

$$\text{Fund.}\begin{cases}0^\circ\\100\%\end{cases} + \text{3rd}\begin{cases}180^\circ\\100\%\end{cases} + \text{5th}\begin{cases}0^\circ \text{ to } -90^\circ\\5\% \text{ to } 100\%\end{cases}$$



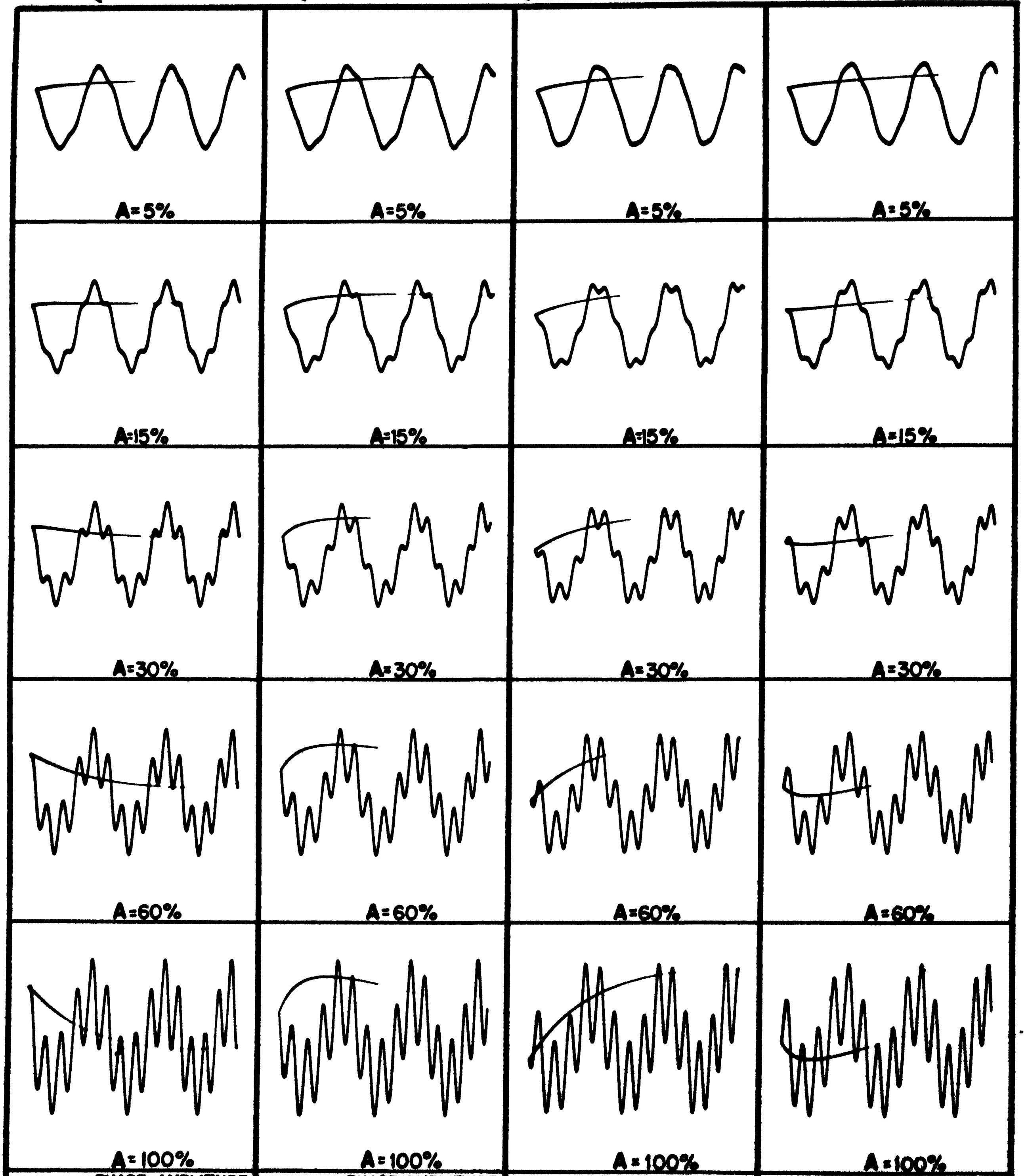
	PHASE	AMPLITUDE
Fund.	0°	100%
3rd	180°	100%
5th	0°	see A

	PHASE	AMPLITUDE
Fund.	0°	100%
3rd	180°	100%
5th	+90°	see A

	PHASE	AMPLITUDE
Fund.	0°	100%
3rd	180°	100%
5th	180°	see A

	PHASE	AMPLITUDE
Fund.	0°	100%
3rd	180°	100%
5th	-90°	see A

$$\text{Fund.} \left\{ \begin{matrix} 0^\circ \\ 100\% \end{matrix} \right. + \text{3rd} \left\{ \begin{matrix} -90^\circ \\ 5\% \end{matrix} \right. + \text{5th} \left\{ \begin{matrix} 0^\circ \text{ to } -90^\circ \\ 5\% \text{ to } 100\% \end{matrix} \right.$$



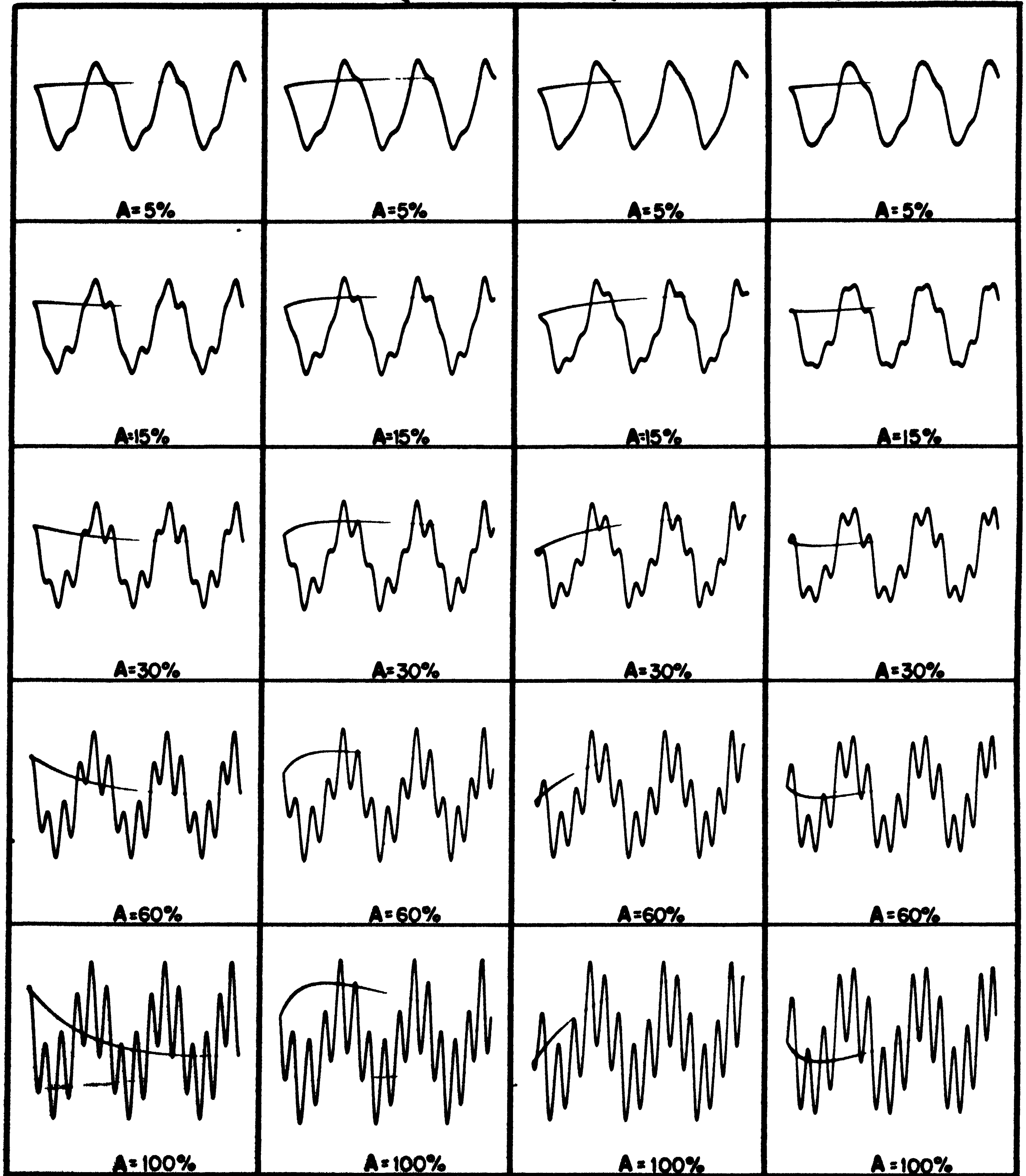
	PHASE	AMPLITUDE
Fund.	0°	100%
3rd	-90°	5%
5th	0°	see A

	PHASE	AMPLITUDE
Fund.	0°	100%
3rd	-90°	5%
5th	+90°	see A

	PHASE	AMPLITUDE
Fund.	0°	100%
3rd	-90°	5%
5th	180°	see A

	PHASE	AMPLITUDE
Fund.	0°	100%
3rd	-90°	5%
5th	-90°	see A

$$\text{Fund.}\begin{cases}0^\circ\\100\%\end{cases}+\ \text{3rd}\begin{cases}-90^\circ\\15\%\end{cases}+\ \text{5th}\begin{cases}0^\circ \ \text{to}\ -90^\circ\\5\% \ \text{to}\ 100\%\end{cases}$$



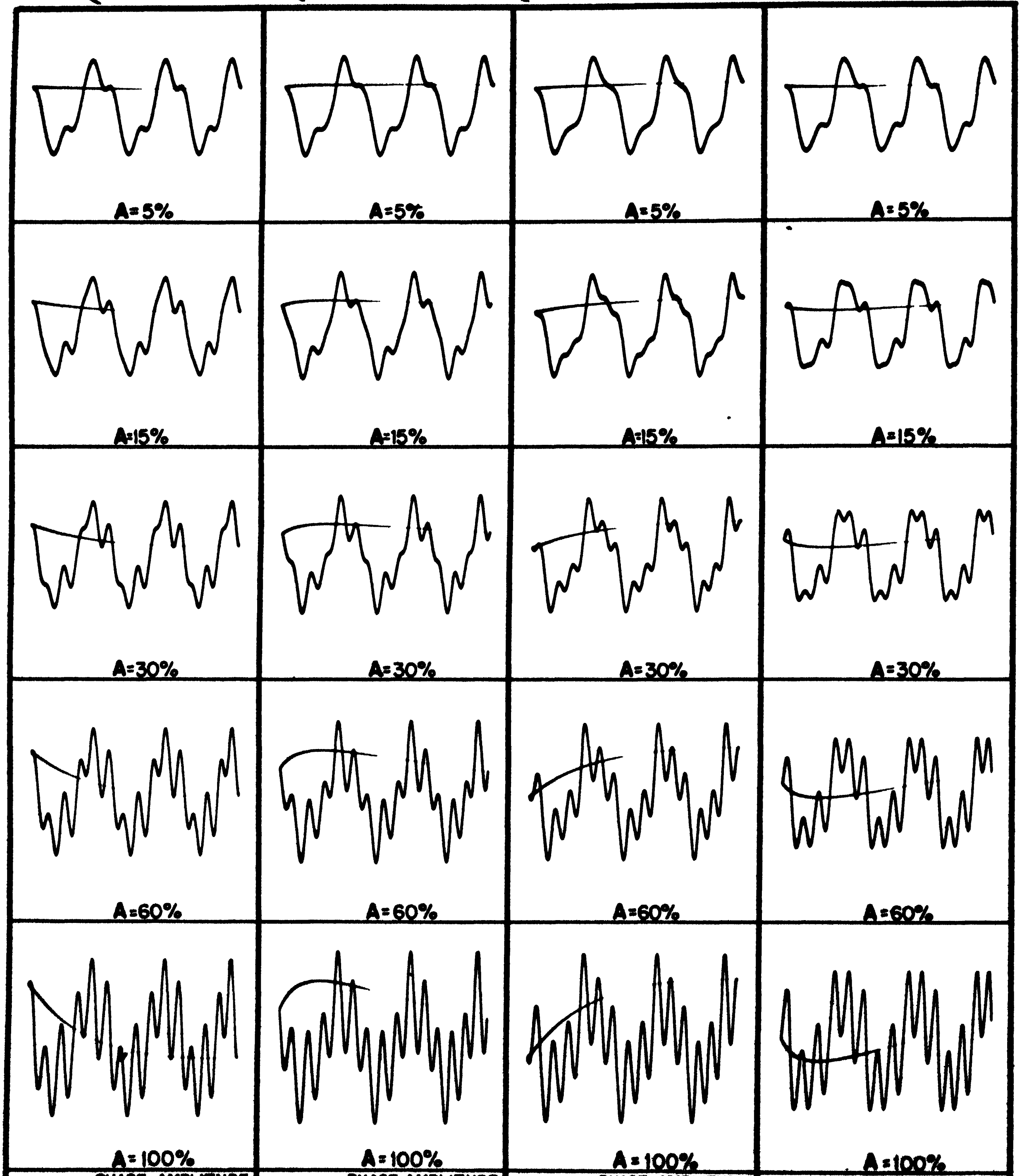
	PHASE	AMPLITUDE
Fund.	0°	100%
3rd	-90°	15%
5th	0°	*see* A

	PHASE	AMPLITUDE
Fund.	0°	100%
3rd	-90°	15%
5th	+90°	*see* A

	PHASE	AMPLITUDE
Fund.	0°	100%
3rd	-90°	15%
5th	180°	*see* A

	PHASE	AMPLITUDE
Fund.	0°	100%
3rd	-90°	15%
5th	-90°	*see* A

$$\text{Fund.}\left\{\begin{matrix}0^\circ\\100\%\end{matrix}\right. + \text{3rd}\left\{\begin{matrix}-90^\circ\\30\%\end{matrix}\right. + \text{5th}\left\{\begin{matrix}0^\circ \text{ to } -90^\circ\\5\% \text{ to } 100\%\end{matrix}\right.$$



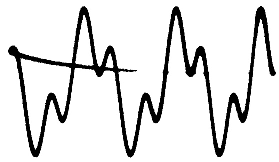
	PHASE	AMPLITUDE
Fund.	0°	100%
3rd	-90°	30%
5th	0°	*see* A

	PHASE	AMPLITUDE
Fund.	0°	100%
3rd	-90°	30%
5th	+90°	*see* A

	PHASE	AMPLITUDE
Fund.	0°	100%
3rd	-90°	30%
5th	180°	*see* A

	PHASE	AMPLITUDE
Fund.	0°	100%
3rd	-90°	30%
5th	-90°	*see* A

$$\text{Fund.}\begin{cases}0^\circ\\100\%\end{cases}+\ \text{3rd}\begin{cases}-90^\circ\\60\%\end{cases}+\ \text{5th}\begin{cases}0^\circ\ \text{to}\ -90^\circ\\5\%\ \text{to}\ 100\%\end{cases}$$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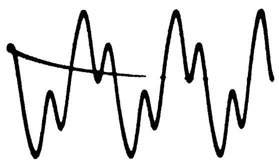
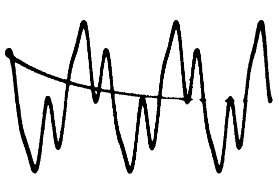

A=15%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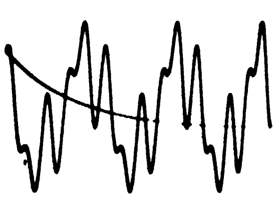
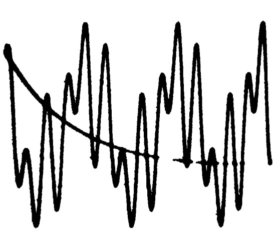
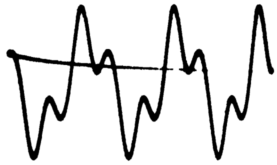

A=60%


A=100%

	PHASE	AMPLITUDE
Fund.	0°	100%
3rd	-90°	60%
5th	0°	see A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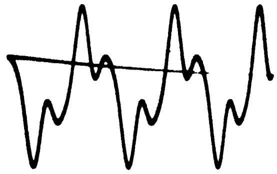
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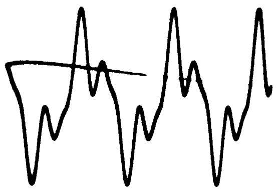
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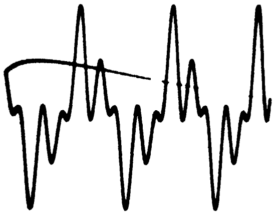
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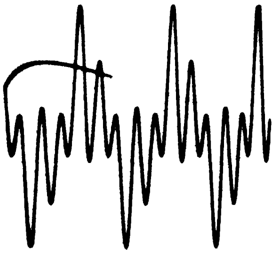
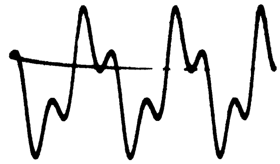

A=100%

	PHASE	AMPLITUDE
Fund.	0°	100%
3rd	-90°	60%
5th	+90°	see A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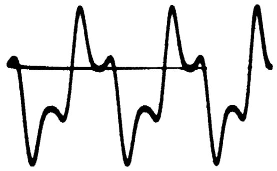
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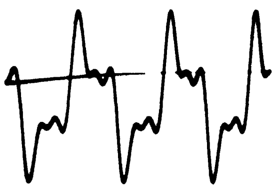
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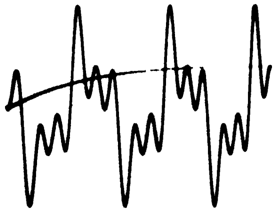
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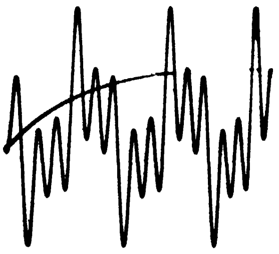
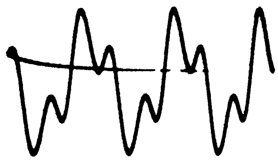
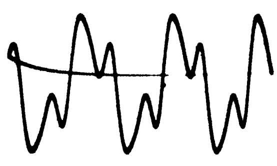

A=100%

	PHASE	AMPLITUDE
Fund.	0°	100%
3rd	-90°	60%
5th	180°	see A


A=5%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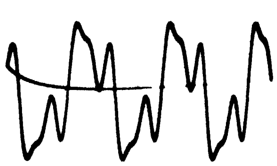
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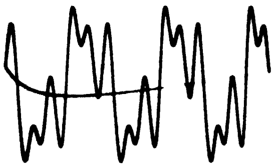
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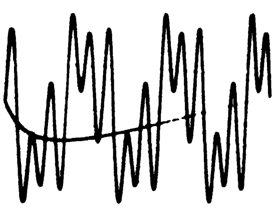
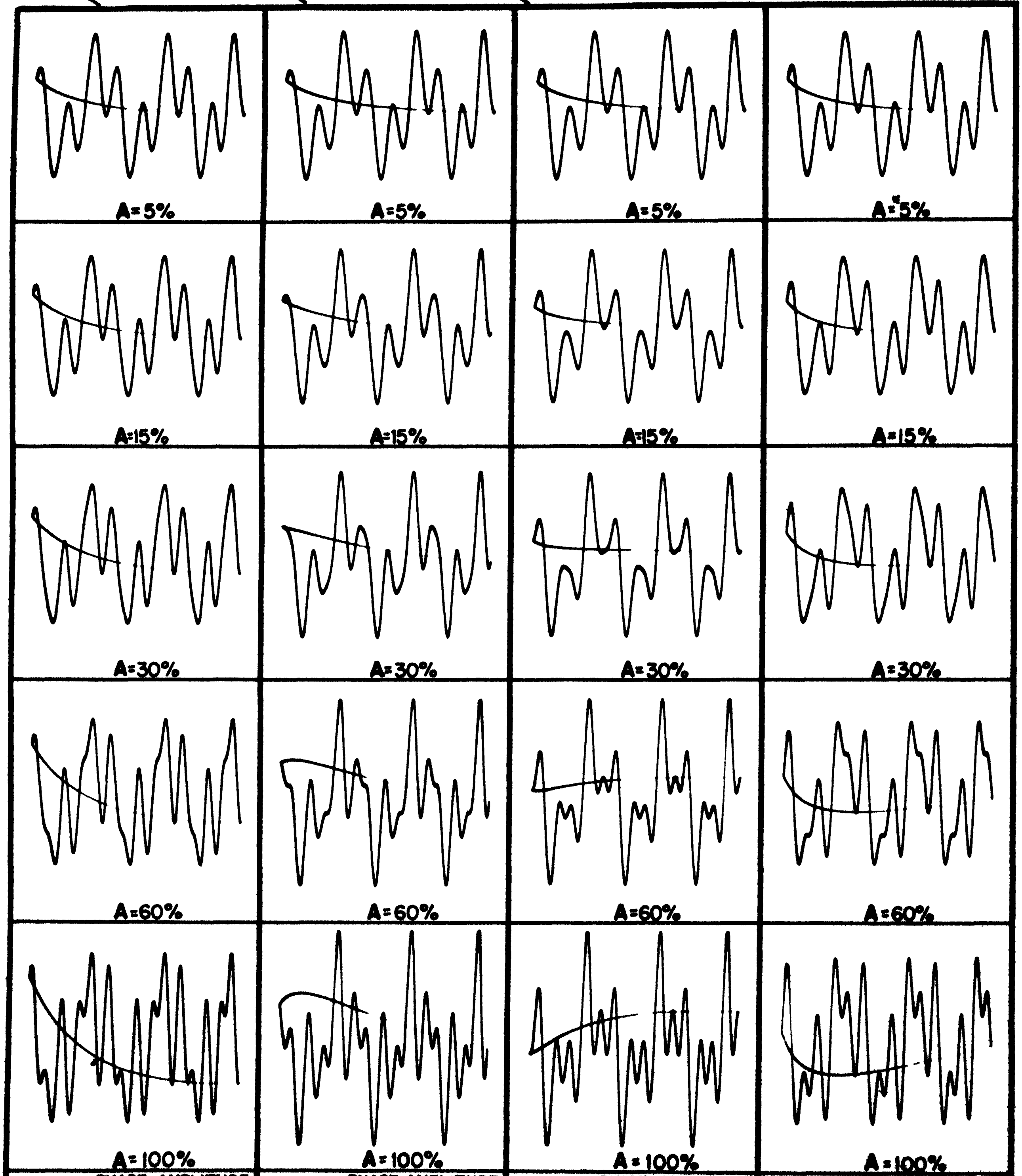

A=100%

	PHASE	AMPLITUDE
Fund.	0°	100%
3rd	-90°	60%
5th	-90°	see A

$$\text{Fund.}\left\{\begin{matrix}0^\circ\\ 100\%\end{matrix}\right. + \text{3rd}\left\{\begin{matrix}-90^\circ\\ 100\%\end{matrix}\right. + \text{5th}\left\{\begin{matrix}0^\circ \text{ to } -90^\circ\\ 5\% \text{ to } 100\%\end{matrix}\right.$$



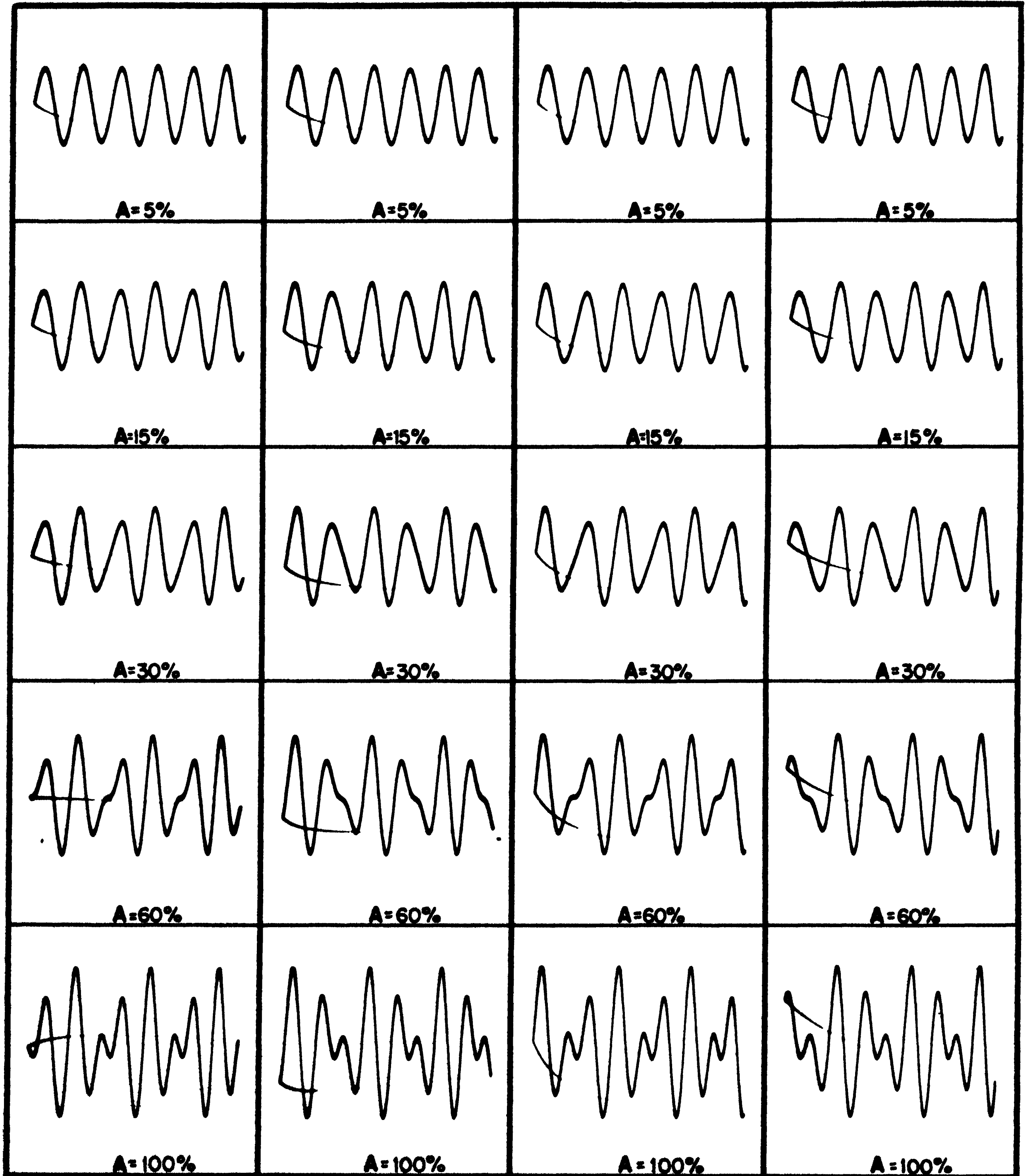
	PHASE	AMPLITUDE
Fund.	0°	100%
3rd	-90°	100%
5th	0°	*see* A

	PHASE	AMPLITUDE
Fund.	0°	100%
3rd	-90°	100%
5th	+90°	*see* A

	PHASE	AMPLITUDE
Fund.	0°	100%
3rd	-90°	100%
5th	180°	*see* A

	PHASE	AMPLITUDE
Fund.	0°	100%
3rd	-90°	100%
5th	-90°	*see* A

2nd $\left\{\begin{matrix}0^\circ\\100\%\end{matrix}\right.$ + 3rd $\left\{\begin{matrix}0^\circ & \text{to} & -90^\circ\\5\% & \text{to} & 100\%\end{matrix}\right.$



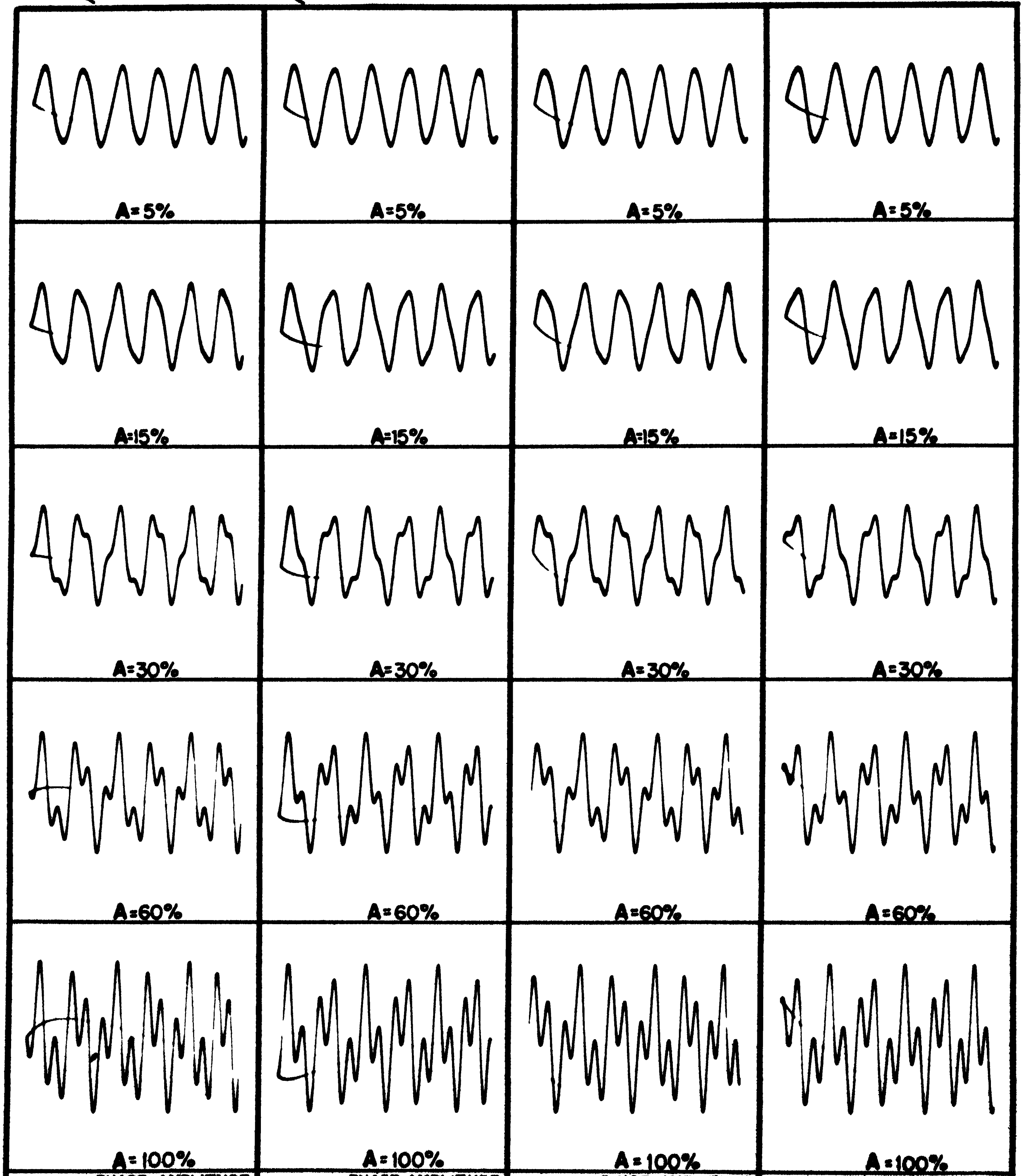
	PHASE	AMPLITUDE
2nd	0°	100%
3rd	0°	see A

	PHASE	AMPLITUDE
2nd	0°	100%
3rd	+90°	see A

	PHASE	AMPLITUDE
2nd	0°	100%
3rd	180°	see A

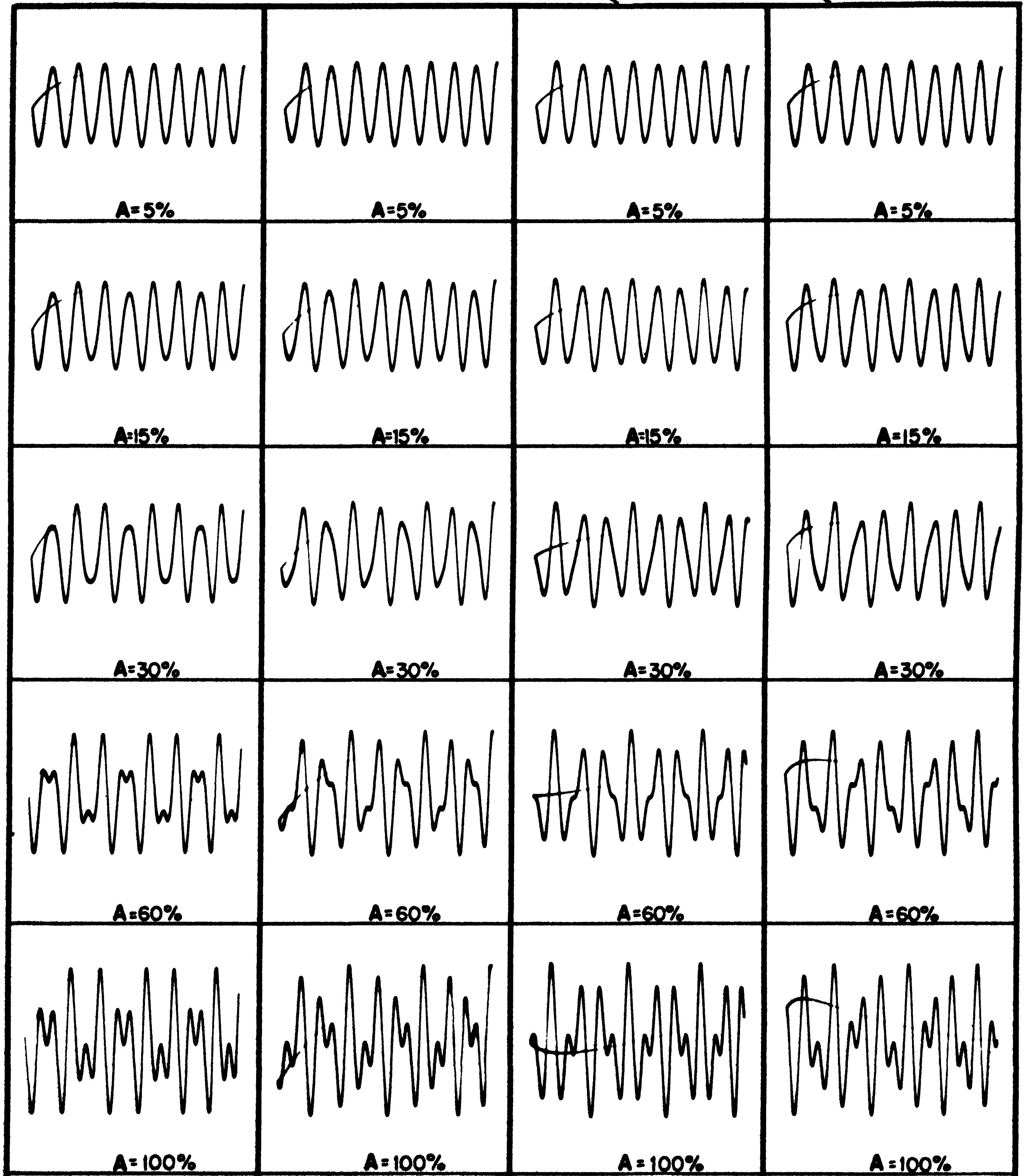
	PHASE	AMPLITUDE
2nd	0°	100%
3rd	-90°	see A

$$2\text{nd}\left\{\begin{matrix}0^\circ\\100\%\end{matrix}\right. + \;5\text{th}\left\{\begin{matrix}0^\circ & \text{to} & -90^\circ\\5\% & \text{to} & 100\%\end{matrix}\right.$$



	PHASE	AMPLITUDE		PHASE	AMPLITUDE		PHASE	AMPLITUDE		PHASE	AMPLITUDE
2nd	0°	100%	2nd	0°	100%	2nd	0°	100%	2nd	0°	100%
5th	0°	*see* A	5th	+90°	*see* A	5th	180°	*see* A	5th	-90°	*see* A

3rd { 0° / 100% } + 5th { 0° to −90° / 5% to 100% }



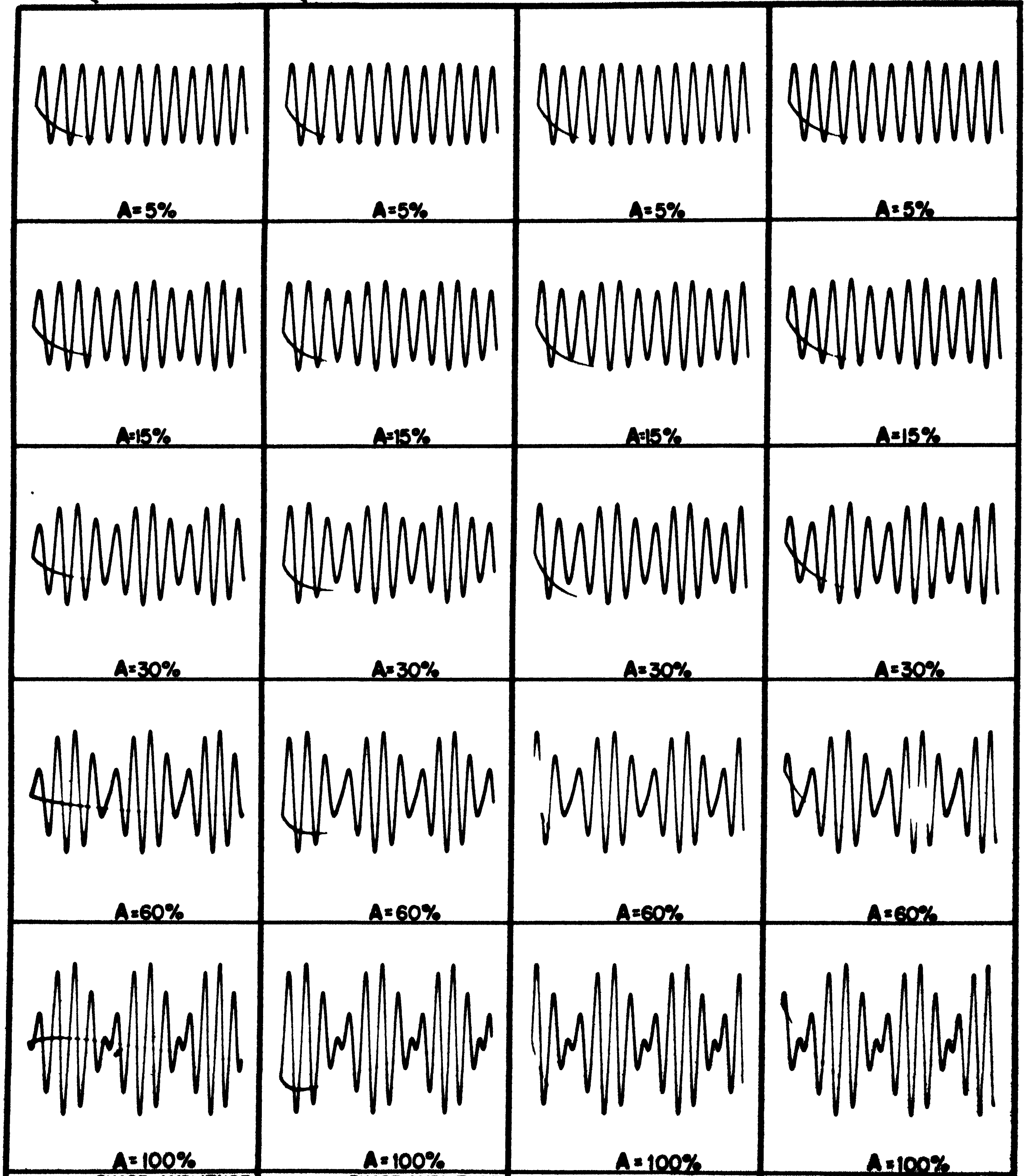
	PHASE	AMPLITUDE
3rd	0°	100%
5th	0°	*see* A

	PHASE	AMPLITUDE
3rd	0°	100%
5th	+90°	*see* A

	PHASE	AMPLITUDE
3rd	0°	100%
5th	180°	*see* A

	PHASE	AMPLITUDE
3rd	0°	100%
5th	-90°	*see* A

$$4\text{th}\begin{cases}0^\circ\\100\%\end{cases} + \;5\text{th}\begin{cases}0^\circ \text{ to } -90^\circ\\5\% \text{ to } 100\%\end{cases}$$



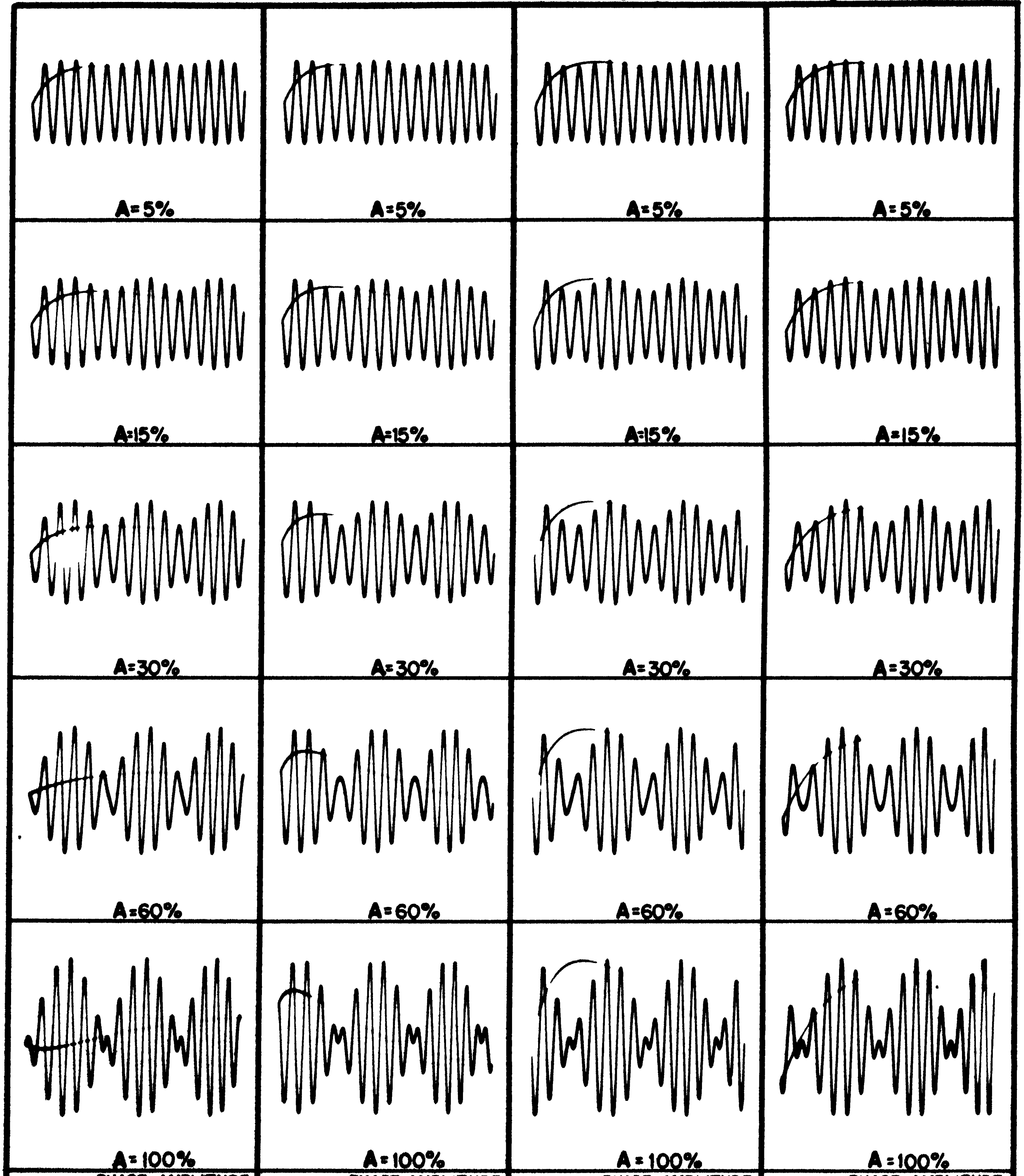
	PHASE	AMPLITUDE
4th	0°	100%
5th	0°	see A

	PHASE	AMPLITUDE
4th	0°	100%
5th	+90°	see A

	PHASE	AMPLITUDE
4th	0°	100%
5th	180°	see A

	PHASE	AMPLITUDE
4th	0°	100%
5th	-90°	see A

5th {0° 100% + 6th {0° to −90° 5% to 100%



	PHASE	AMPLITUDE
5th	0°	100%
6th	0°	*see A*

	PHASE	AMPLITUDE
5th	0°	100%
6th	+90°	*see A*

	PHASE	AMPLITUDE
5th	0°	100%
6th	180°	*see A*

	PHASE	AMPLITUDE
5th	0°	100%
6th	-90°	*see A*

CHAPTER 21

SPECIAL PURPOSE CATHODE-RAY TUBES

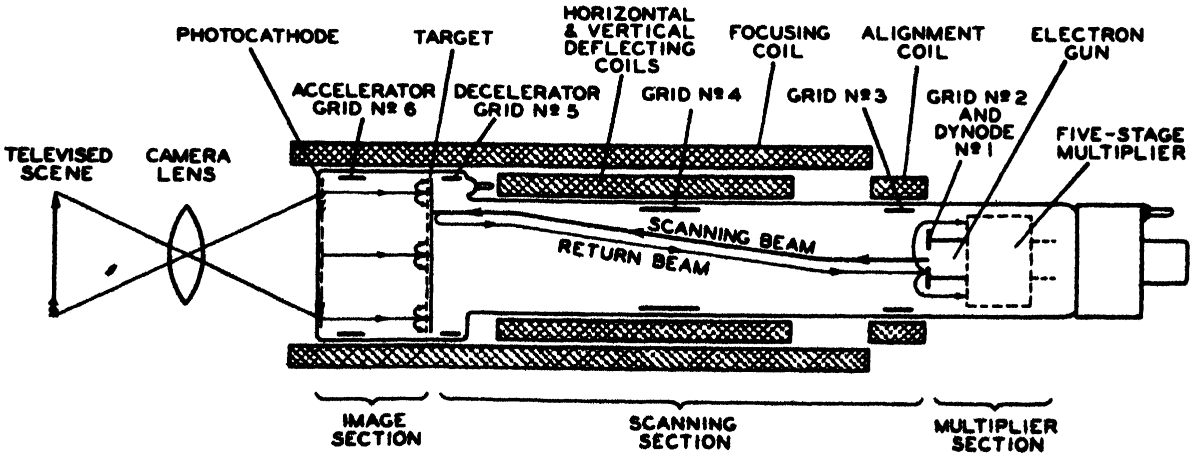
This chapter will deal with special cathode-ray tubes as distinguished from conventional types used in oscilloscopes and television receivers. The sequence of discussion will be in order of relative importance. The image orthicon, widely used in television broadcasting pickups, will be discussed first. The image iconoscope, while still used to some extent, appears to be growing obsolete. The image dissector, also once used extensively, has assumed less importance because of its comparatively low sensitivity but is still used in specialized applications. The monoscope, transcriber kinescope, flying-spot cathode-ray tube, and graphecon are other types that will be discussed.

Image Orthicon[1]

This tube is shown, in schematic diagram form, in Fig. 21-1. For simplicity in discussing it, three sections may be studied: the image section, scanning section, and multiplier section. A photograph of a typical type appears in Fig. 21-2.

Courtesy RCA

Fig. 21-2.—Photograph of an image orthicon.



Courtesy RCA

Fig. 21-1.—Schematic diagram of the image orthicon.

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[1] The following material on the image orthicon consists of excerpts from the RCA Tube Data Sheets, 5655 and 5679.

Image Section

The image section contains a semitransparent photocathode on the inside of the face plate, a grid to provide an electrostatic accelerating field, and a target which consists of a thin glass disk with a fine mesh screen very closely spaced to it on the photocathode side. Focusing is accomplished by means of a magnetic field produced by an external coil, and by varying the photocathode voltage.

Light from the scene being televised is picked up by an optical lens system and focused on the photocathode which emits electrons from each illuminated area in proportion to the intensity of the light striking the area. The streams of electrons are focused on the target by the magnetic and accelerating fields.

On striking the target, the electrons cause secondary electrons to be emitted by the glass. The secondaries thus emitted are collected by the adjacent mesh screen which is held at a definite potential of about 2 volts with respect to target voltage cutoff. Therefore, the potential of the glass disk is limited for all values of light and stable operation is achieved. Emission of the secondaries leaves, on the photocathode side of the glass, a pattern of positive charges which corresponds with the pattern of light from the scene being televised. Because of the thinness of the glass, the charges set up a similar potential pattern on the opposite or scanned side of the glass.

Scanning Section

The opposite side of the glass is scanned by a low-velocity electron beam produced by the electron gun in the scanning section. This gun contains a thermionic cathode, a control grid (grid No. 1), and an accelerating grid (grid No. 2). The beam is focused at the target by the magnetic field of an external focusing coil and the electrostatic field of grid No. 4.

Grid No. 5 serves to adjust the shape of the decelerating field between grid No. 4 and the target to obtain uniform distribution of electrons over the entire target area. The electrons normally stop their forward motion at the surface of the glass and, after being turned back, they are focused into a five-stage signal multiplier. However, when electrons approach the positively charged portions of the pattern on the glass, they are deposited from the scanning beam in quantities sufficient to neutralize the potential pattern on the glass.

This deposition of electrons leaves the glass with a negative charge on the scanned side and a positive charge on the photocathode side. These charges will neutralize each other by conduction through the glass in less time than that taken up by one frame.

Alignment of the beam from the gun is accomplished by a transverse magnetic field produced by an external coil located in the vicinity of the gun at the end of the focusing coil. Deflection of the beam is achieved by transverse magnetic fields produced by external deflection coils.

The electrons turned back at the target form the return beam which has been amplitude-modulated by absorption of electrons at the target in accordance with the charge pattern whose more positive areas correspond to the highlights of the televised scene.

Multiplier Section

The return beam is directed to the first "dynode" of a five-stage electrostatically focused multiplier. Use is made of the phenomenon of secondary emission to amplify signals composed of electron beams. The electrons in the beam, impinging on the first-dynode surface, produce many other electrons, the number depending on the energy of the impinging electrons. A dynode is a specially treated anode which produces a number of secondary electrons greater than the number of primary electrons impinging on its surface. These secondary electrons are then directed to the second dynode and knock out more new electrons. Grid No. 3 facilitates a more complete collection by dynode No. 2 of the secondaries from dynode No. 1.

The multiplying process is repeated in each successive stage, with an ever-increasing stream of electrons, until those emitted from dynode No. 5 are collected by the anode and constitute the current utilized in the output circuit.

The multiplier section amplifies the modulated beam about 500 times. The multiplication so obtained increases the signal-to-noise ratio of the tube and also permits the use of an amplifier with fewer stages. The gain of the multiplier is sufficiently high so that the limiting noise in the use of the tube is the random noise of the electron beam multiplied by the multiplier stages. This noise is larger than the input noise of the video amplifier.

Iconoscope[2]

The image orthicon is a relatively new development that was preceded by the iconoscope, which is simpler in design and construction, but has a lower sensitivity.

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[2] Zworykin, V. K., "Iconoscopes and kinescopes in television," *RCA Rev.*, vol. 1, pp. 60-84, July 1936. The following material on iconoscopes consists of excerpts from this article.

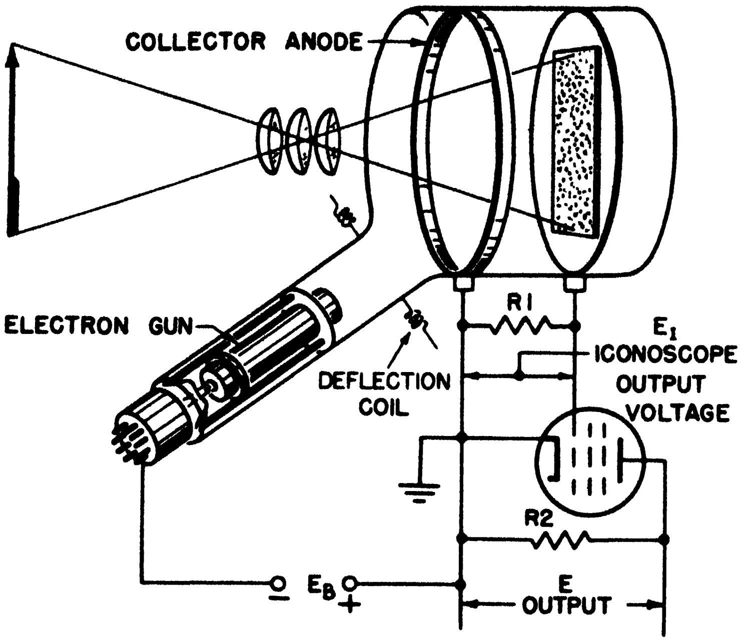


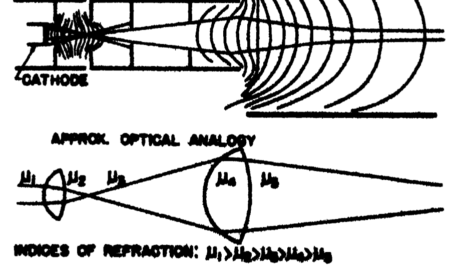
Fig. 21-3.—Iconoscope tube and its circuit.

An iconoscope tube and its circuit are shown in Fig. 21-3. It is a special form of cathode-ray tube used in television for "picking up a scene" and converting it to an electrical signal. The tube contains a mosaic plate on which the scene to be transmitted is focused by a lens, an electron gun which provides a cathode-ray beam for scanning the image on the mosaic, and a signal plate which acquires a voltage proportional to the image brightness of the particular spot being scanned at any instant. This voltage is the "video" signal for the television picture.

Constructional Details

The principal parts of an iconoscope are the mosaic, signal plate, collector, and electron gun. The tube is illustrated in Fig. 21-3. The electron gun produces a narrow pencil of cathode-rays which serves, as will be shown later, as a commutator to the tiny photocells on the mosaic. The gun is in reality a form of electron projector which concentrates the electrons from the cathode onto the mosaic in a very small spot. The electron-optical system consists of two electron lenses which are formed by the cylindrically symmetrical electrostatic fields between the elements of the gun. Fig. 21-4 shows diagrammatically the arrangement of the gun, together with the electrostatic fields making up the electron lenses. Below this diagram is the approximate optical analogue. Details of the gun construction are as follows: The cathode is indirectly heated with its emitting area at the tip of the cathode cylinder. It is mounted so that the emitting area is a few thousandths of an inch in front of an aperture in the control grid.

A long cylinder with three defining apertures whose axis coincides with that of the cathode cylinder and control grid serves to give the electrons their initial acceleration and is known as the first anode. A second cylinder coaxial with the first anode and of somewhat greater diameter serves as the second anode and gives the electrons their final velocity. The second anode is, in general, formed by metalizing the neck of the iconoscope bulb. The gun used in the iconoscope is designed so that it will concentrate a beam current of from ½ to 1 microampere into a spot about 5 mils in diameter. Under ordinary operating conditions, a potential of about 1,000 volts is applied between the cathode and second anode and the voltage of the first anode adjusted until minimum spot size is obtained. The exact



Courtesy RCA Rev.

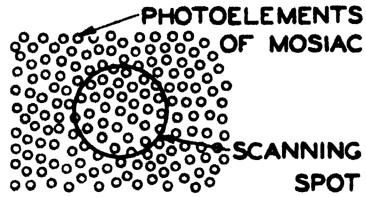
Fig. 21-4.—Diagram of the electron-optical system of the iconoscope.

value of the beam current to be used will, of course, depend upon the type of picture to be transmitted and the exact conditions of operation.

The beam from the gun is made to scan the mosaic in a series of parallel horizontal lines repeated at 30 cps. This is accomplished by two sets of magnetic deflection coils arranged in a suitable yoke and slipped over the neck of the tube. These sets of coils are driven by two special vacuum-tube generators supplying a sawtoothed current wave; one, operating at the picture frequency, supplies the vertical-deflection coils, the other at horizontal line frequency drives the second set of coils.

The component which characterizes the iconoscope is the mosaic. It consists of a vast number of photosensitive globules mounted on a thin mica sheet in such a way that they are insulated from one another. The back of this sheet is coated with a conducting metallic film which serves as a signal plate and is connected to the input of the picture amplifier. The appearance of the mosaic is shown in Fig. 21-5. Such a mosaic may be formed in a variety of ways. For the standard type of mosaic, the silver globules are formed by reducing particles of silver oxide dusted over the mica. Under proper heat treatment, the silver globules reduced from the oxide will not coalesce but will form individual droplets. These droplets are sensitized after the mosaic has been mounted in the tube and the tube evacuated. The sensitization is similar to that used in the ordinary cesium photocell; that is, the silver is oxidized, exposed to cesium vapor, and then heat-treated.

Fig. 21-5.—The mosaic of the iconoscope, greatly magnified.



Courtesy RCA Rev.

Photoelectric Response

The mica on which the silver droplets are mounted serves to insulate them from one another; it is made sufficiently thin so that the capacitance between each globule and the metallic signal plate will be reasonably large.

The mosaic is mounted in the tube with the silver beads facing the beam. For proper focusing of the optical image on the front surface of the iconoscope, it is placed in the tube in such a way that a normal to its face makes an angle of 30° to the axis of the electron gun.

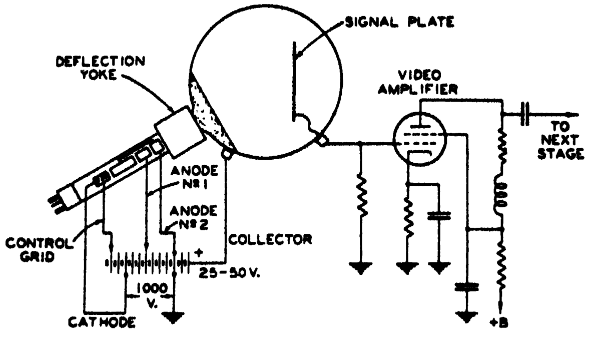
In essence, the iconoscope may be thought of as a plain mosaic made up of a great number of individual photocells, all connected capacitively to the common signal plate and commutated by the scanning beam. The fundamental cycle of operation is as follows:

Every silver globule making up the mosaic is photosensitized so that when a light image is projected on the mosaic the light causes electrons of a number proportional to the light brilliance to be emitted from each illuminated photosensitive area. The resulting loss of electrons leaves each photosensitive area at a positive potential with respect to its initial condition. The potential is then proportional to the number of electrons which have been released and conducted away so that the mosaic tends to go positive at a rate proportional to the light falling on it. As the electron beam scans the mosaic, it passes over each element in turn, releasing the charge it has acquired and driving it to equilibrium.

Due to the fact that each element is coupled capacitively to the signal plate, the sudden change of charge of the elements will induce a change in charge on the signal plate resulting in a current pulse in the signal lead connected to the amplifier. The magnitude of these pulses is proportional to the intensity of the light falling on the scanned element. Thus the signal output from the iconoscope consists of a chain of current pulses corresponding to the light distribution on the mosaic. This chain can be resynthesized at the receiver into a reproduction of the original image.

As the beam moves over the mosaic, varying amounts of charge flow from the mosaic to the collector, the amount of charge flowing at any instant being a measure of the light on the globules at the instantaneous position of the beam. In other words, a video signal current flows between the mosaic and collector. It can be seen that the video signal current must complete its circuit path through the signal-plate equivalent load resistor *R1* and the capacitance between signal plate and mosaic. The voltage developed across this equivalent load resistor by the signal current is the video signal output of the iconoscope.

It also can be seen that when the beam moves from a dark portion of the image to a brighter portion, the electron current from the mosaic to the collector decreases. The output voltage, therefore, changes in the negative direction. Hence the signal output of the iconoscope is of negative polarity. A highlight in the image is represented by a negative value of signal voltage; a shadow in the image is represented by a positive value.

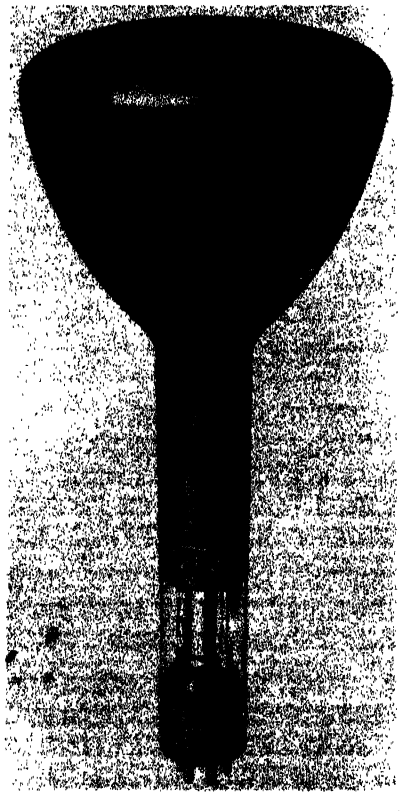


Courtesy RCA Rev.

Fig. 21-6.—The monoscope and its circuit.

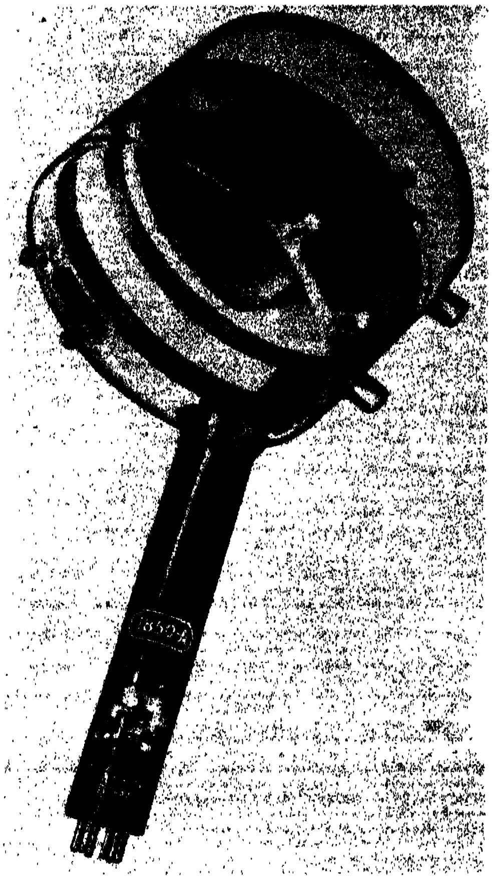
Monoscope[8]

"Monoscope" is the name which has been given to a type of tube designed to produce a video signal of a test picture or pattern enclosed in the tube. Since the picture must be enclosed in the tube, the monoscope is not suitable for developing a video signal which represents action. However, a signal of excellent fidelity can be obtained for a "still" picture which contains half-tones or consists only of lines. Fig. 21-6 shows the circuit used. Two examples of such tubes are shown in Figs. 21-7 and 21-8. The former is the RCA 2F21-A and the latter is the magnetically deflected RCA 1850-A.



Courtesy RCA

Fig. 21-7.—Electrostatic monoscope, RCA type 2F21.



Courtesy RCA

Fig. 21-8.—Electromagnetic monoscope, RCA type 1850-A.

The tube consists of an electron gun, a signal plate, and a collector enclosed in a highly evacuated envelope. The electron beam is scanned over the signal plate by an electromagnetic-deflection system.

The electron gun which supplies the scanning beam must be of high quality if the best video signal is to be obtained. The electron beam should be very small in diameter, when it strikes the signal plate, if good resolution is to be obtained. The beam current should be reasonably high because the video current varies directly with beam current. However, the width of the beam should not be sacrificed to obtain high beam currents. The electron gun developed for iconoscopes provides a small diameter beam with an adequate beam current and, therefore, was readily adapted to the monoscope.

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[8] Burnett, C. E., "The monoscope," *RCA Rev.*, vol. 2, pp. 414-420, April 1938. The material on the monoscope consists of excerpts from this article.

The final anode of this gun operates at 1,000 volts, permitting a very small beam for currents of several microamperes. The use of this gun proves to be an advantage in test work when it is desired to use a monoscope in an iconoscope camera.

The signal plate is made of aluminum foil and carbon. The surface of the aluminum has a natural coating of aluminum oxide which has a reasonably high secondary-emission ratio while the carbon has a relatively low ratio. It was found that aluminum foil developed for advertising and packing purposes, as well as special inks developed for printing on metal foils, were satisfactory materials for signal plates. As a result, the advantages and flexibility of commercial printing processes can be utilized. The desired picture or pattern is printed on aluminum foil with a black foil ink. The only other processing necessary before sealing the signal plate in the tube is to fire it in hydrogen. This process removes the volatile matter from the ink and thus leaves it practically pure carbon.

Correct Polarity

In order to give the picture the correct polarity on the kinescope; i.e., so that white corresponds to white in the original, it is necessary to make the signal plate in a definite manner, depending on the number of stages in the video amplifier. If the video amplifier has an odd number of stages (as is normal between iconoscope and kinescope) the picture on the signal plate of the monoscope should have blacks and whites reversed, but should not have printed matter reversed. The reversal of blacks and whites is necessary because the aluminum oxide, although white in appearance, has a higher secondary-emission ratio than the carbon, and, therefore, produces a signal which corresponds to black.

The secondary-emission current from the signal plate is collected by a conductive coating on the bulb wall. This coating is operated at a potential positive with respect to the signal plate, which is operated at the same potential as the final anode of the electron gun.

Uses

The monoscope may be used for a variety of purposes. In commercial applications, frequently repeated announcements and advertisements could be taken from a monoscope. Fixed backgrounds for studio work could be obtained; the final signal being a suitable combination of video signal from an iconoscope for action and from a monoscope for background. However, the biggest field for the monoscope is in television testing. The same video signal can be obtained from day to day, and the quality is not affected by such variables as poor optical focus, dark spot, and amplifier noise. The monoscope, therefore, is an important device for checking performance of kinescopes, receivers, and studio systems.

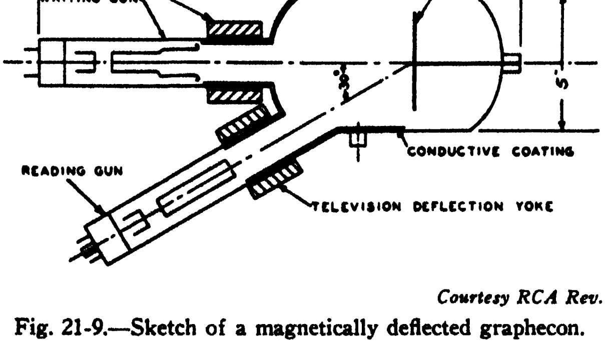
Graphecon[4]

Long-time storage of television quality pictures is possible with a type of storage tube consisting of two independent cathode-ray guns and a target plate coated with a thin film of insulating material. The two guns may be operated simultaneously. One "writes" down any arbitrary pattern to be stored. The other scans repeatedly over it to generate signals and erase the pattern at a controlled rate. The graphecon also makes possible television pictures of oscillograms or radar patterns that can be viewed for several minutes.

The tube can be regarded as a kinescope and iconoscope in one bulb, with the mosaic and kinescope screen replaced by a charge-sensitive, high-capacitance storage target. The kinescope gun will be referred to as the "writing" gun which takes the radar signal and "writes" it on the target. The other gun will be called a "reading" gun. Its function is to generate the television signal and gradually remove the stored signals.

Construction

The design of the bulb is not critical and is largely determined by the requirements of the cathode-ray guns. Fig. 21-9 shows a sketch of one type of tube using magnetic deflection. To simplify the circuit re-



Courtesy RCA Rev.

Fig. 21-9.—Sketch of a magnetically deflected graphecon.

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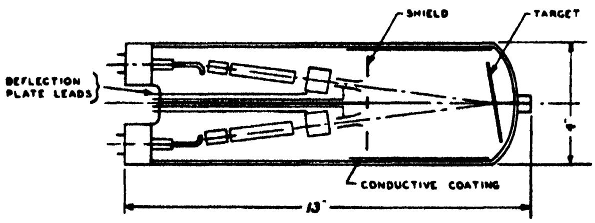
[4] Pensak, L., "The graphecon—a picture storage tube," *RCA Rev.*, vol. 10, pp. 59-73, March 1949; the material on the graphecon in the following section was abstracted from this article.

quirements, the writing gun is mounted perpendicular to the target, which avoids the necessity for keystone correction for a radial deflection pattern. This correction has been solved very simply for the iconoscope, and so the reading gun is mounted off the axis, at the same angle as is used for the standard iconoscope. The dimensions of the bulb are determined by the space requirements of the deflection yokes, by the deflection angles required for obtaining adequate resolution, and by the desired target dimensions.

Focusing

When preferable, a tube can be built with all-electrostatic focus and deflection. It is more suitable for oscilloscope operation or, where weight and size are critical, it permits eliminating the deflection yokes. Fig. 21-10. is a sketch of such a tube which can be made smaller and more compact than the magnetic-deflection type, but is necessarily subject to the lower resolution limits imposed by the electrostatic-deflection system.



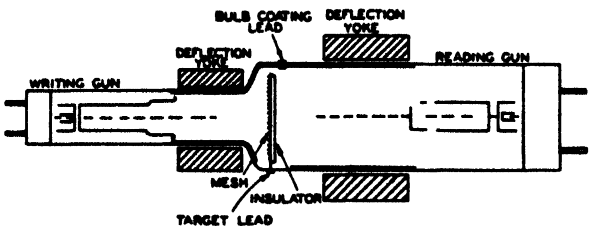
Courtesy RCA Rev.

Fig. 21-10.—Sketch of a compact graphecon utilizing electrostatic focusing and deflection.

Fig. 21-11 is a sketch of the most recent form of the tube. This modification was built to make possible mounting the guns on a common axis and thereby avoid the need for keystone correction. The target construction was modified to permit the "writing" beam to penetrate through the target.

For the magnetic-deflection writing gun, it is possible to use a standard kinescope gun with either magnetic focus or electrostatic focus. Both types have been used for radar purposes and have adequate resolution at the rated voltages of 6,000 to 10,000 volts, which is also the range of voltage which is used in the graphecon. The reading gun can be a standard iconoscope gun and runs at standard iconoscope voltages of 800 to 1,000 volts.

The target consists of a plate of metal upon which is deposited a film of insulating material of the order of 6,000 Angstrom units thick, or approximately half a micron. Where the two guns are on the same side, any of a number of metals, thick enough to be self-supporting, can be used for the target. The insulator film can be any good insulator such as silica or magnesium fluoride.



Courtesy RCA Rev.

Fig. 21-11.—Sketch of the most recent form of the graphecon.

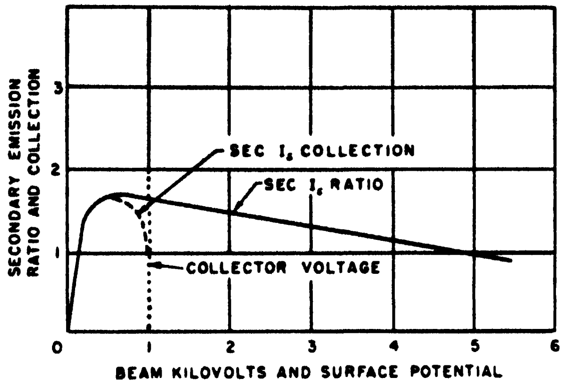
In tubes where the guns are on opposite sides of the target, it is necessary to make the metal backing transparent to electrons and yet strong enough to support the insulating layer. This can be done by using a very fine mesh with a high transmission factor to provide the mechanical support. Then, an organic film is spread over the mesh to act as a base upon which to evaporate a thin layer of aluminum. The insulating layer is then evaporated onto the aluminum and the target is completed. The high-voltage "writing" gun is located on the mesh side of the target. The mesh is made fine enough so that it does not limit the resolving power of the tube. Sample targets have employed approximately 500 per-inch mesh.

Operation

The reading beam, operating at 1,000 volts, has a secondary-emission ratio greater than unity. It scans uniformly over the insulator surface and, therefore, brings it approximately to the potential of the collector which is the conductive wall coating.

The mechanism for signal generation is a simple form of that used in the iconoscope. The target surface is brought to equilibrium potential with the target metal. Where the writing beam has struck and driven the surface negative, the secondary-emission collection is greater than unity and some charge is removed every time the reading beam scans those areas. This removal of charge produces the signal which is amplified to operate the viewing kinescope.

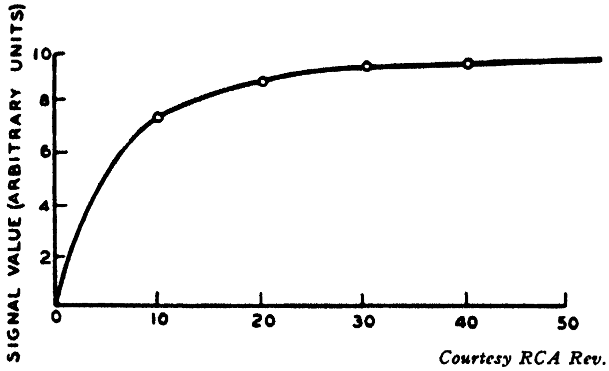
The magnitude of the signal depends on the deviation from the equilibrium condition. When the surface is at collector potential, the secondary-emission collection is equal to the beam current and no signal results. The dotted line in Fig. 21-12 shows how the secondary-emission collection approaches the trace secondary-emission curve when the surface potential falls below



Courtesy RCA Rev.

Fig. 21-12.—Typical curve for secondary emission from silica.

the collector potential. Fig. 21-13 shows this in terms of signal value for a typical electrode configuration. The curve levels off at the higher-voltage values because the secondary-emission collection tends to saturate at relatively weak fields, and the signal current, therefore, remains almost constant at the higher fields.



Courtesy RCA Rev.

Fig. 21-13.—Surface potential in volts below collector potential.

The saturation of secondary emission provides a means of obtaining many television pictures of the writing pattern. The saturated secondary emission is proportional to the beam current so that it is only necessary to reduce the beam current to reduce the amount of charge on each scan. However, this also reduces the signal output and the lower limit in this direction is the noise inherent in the video amplifier.

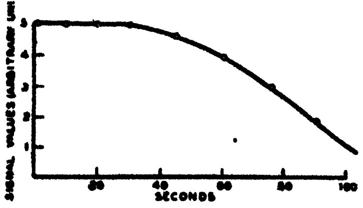


Fig. 21-14.—Typical signal output plotted against viewing time.

Courtesy RCA Rev.

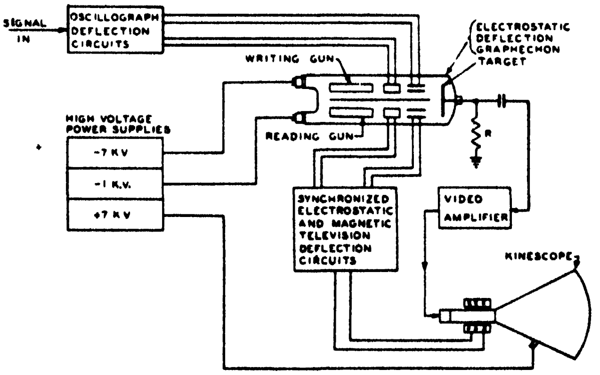
The charging time is proportional to the total area of the target. A reduction in tube size will result, therefore, in a proportional reduction in the maximum viewing time. A typical illustrative curve is shown in Fig. 21-14.

Practice

One of the applications of the tube is as a direct-current or single-trace oscilloscope which can provide a bright picture of the complete trace for adjustable periods of time.

The circuits for such a function are relatively complex, compared to an oscilloscope with an afterglow-type phosphor, but there are several advantages that may justify the extra equipment. The problem of large-screen oscillography can be met by using a commercial television projection set and operating the graphecon as an oscilloscope-iconoscope combination. The problem of photographing transients of very short duration can be met by recording the trace in the graphecon and photographing the picture on a standard kinescope. If the interval between transients is long, the reading beam and kinescope need only be turned on, when desired, for photographing. In either case, the tube can be made very sensitive to the writing beam and thus eliminates the need for a very high-voltage type oscilloscope which ordinarily might be required for high writing speeds.

For such applications, where the writing beam is not modulated in intensity, the circuit requirements are shown in block form in Fig. 21-15.


SIGNAL IN
OSCILLOGRAPH DEFLECTION CIRCUITS
WRITING GUN
ELECTROSTATIC DEFLECTION
GRAPHECHON TARGET
HIGH VOLTAGE POWER SUPPLIES
-7 KV
-1 K.V.
+7 KV
READING GUN
R
SYNCHRONIZED ELECTROSTATIC AND MAGNETIC TELEVISION DEFLECTION CIRCUITS
VIDEO AMPLIFIER
KINESCOPE

Courtesy RCA Rev.

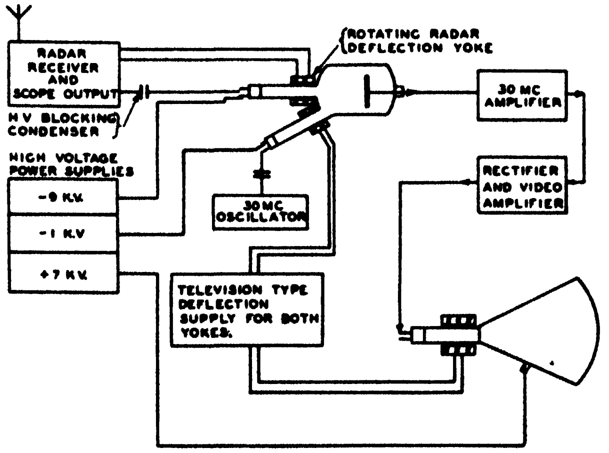
Fig. 21-15.—Block diagram of circuits using graphecon as an oscilloscope.

The requirement that the writing beam not be modulated in intensity arises from the fact that it will produce a video signal in the target of the same form as the modulation and of an intensity that could be many times that of the signal produced by the reading beam.

This occurs because of the fact that any removal or addition of charge to the target produces a signal, and the writing beam, being above the second cross-over point, tends to put down charge which will vary in amount with the intensity of the beam. As long as the writing-beam intensity is constant, it produces only a direct-current type of signal which is blocked by the input capacitor of the video amplifier.

However, for many applications, such as for radar conversion, it becomes necessary to modulate the writing beam at frequencies in the same range as that of the video output signal, and it is, therefore, not possible to separate the two signals on the basis of signal frequency.

It is possible to create a frequency difference between the reading and writing signals by modulating the reading beam at some frequency well above the maximum contained in the writing signal. The output signal will now be an amplitude-modulated high-frequency carrier, which can be amplified by conventional means and rectified to provide the desired video signal free of the writing modulation. A typical circuit is shown in block-diagram form in Fig. 21-16, giving the components required to operate the tube as a radar converter.


ROTATING RADAR DEFLECTION YOKE
RADAR RECEIVER AND SCOPE OUTPUT
H V BLOCKING CONDENSER
HIGH VOLTAGE POWER SUPPLIES
-9 K.V.
-1 K.V
+7 K.V.
30 MC AMPLIFIER
RECTIFIER AND VIDEO AMPLIFIER
30 MC OSCILLATOR
TELEVISION TYPE DEFLECTION SUPPLY FOR BOTH YOKES.

Courtesy RCA Rev.

Fig. 21-16.—Block diagram of circuits required for using graphecon to view PPI type radar presentation.

The original objective and main application of this tube is to provide a means of viewing radar PPI patterns. A number of advantages can be attained over conventional practice with systems using tubes made with the P7 type phosphor.

Resolution

The resolution of the stored picture is best discussed in terms of television practice. Because the picture is formed by the action of a moving cathode-ray beam, the smallest discrete picture element possible is the size of the focused spot. Therefore, the maximum picture content possible is the number of elements or adjacent spot areas that can cover the total area of the target. This is true regardless of the type of scanning used. A television-type scan is the most efficient because it arranges the elements as close together as possible without producing any overlapping.

Writing speed is customarily measured, in oscilloscopic practice, in feet or meters per second. So measured, the writing speed of the graphecon is considerably better than 4,000 feet per second. Expressed in more meaningful terms, this is equivalent to better than 9,000,000 elements per second. This speed has been demonstrated by the test of television operation in which a picture containing 300,000 elements was written onto the target in 1/30 of a second. Still higher writing speeds can be obtained, though only at the expense of resolution, because they require higher beam currents and this causes the spot size to increase, everything else held constant.

In addition to the tubes described, there are a number of others of secondary importance: the image dissector, orthicon, transcriber kinescope, and flying-spot cathode-ray tube.

Image Dissector

This tube is illustrated in Fig. 21-17. The essential parts are the photocathode, focus coil, deflection coil, electron multiplier, and the associated optical system.

The photocathode converts the incident light of the optical image into an electrical signal.

The number of electrons emitted at any point is directly proportional to the radiant light flux at the point. As a result, the cathode surface has an electron emission distribution pattern that correlates with the light

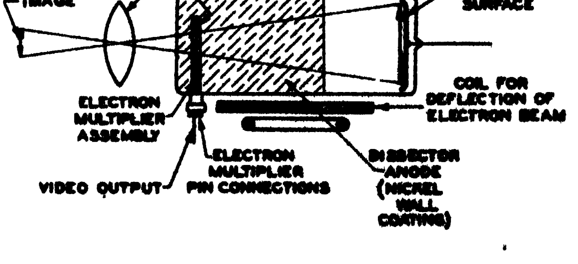

APERTURE FOR SCANNING
FOCUS COIL
OPTICAL LENS
LIGHT IMAGE
PHOTO-SENSITIVE CATHODE SURFACE
COIL FOR DEFLECTION OF ELECTRON BEAM
ELECTRON MULTIPLIER ASSEMBLY
ELECTRON MULTIPLIER PIN CONNECTIONS
DISSECTOR ANODE (NICKEL WALL COATING)
VIDEO OUTPUT

Fig. 21-17.—Farnsworth image dissector tube.

distribution of the optical image and of the scene being televised.

This electron pattern is extended from the cathode surface in straight lines perpendicular to it and is moved toward the aperture in the multiplier housing by a difference of potential between the cathode and the nickel-wall-coating anode. This voltage may be of the order of 350 volts.

As charges of identical polarity repel each other, there is a tendency for the beam electrons to scatter and destroy the image pattern. As the beam moves toward the electron multiplier, the scattering is controlled by means of a focus coil. This coil surrounds the tube to permit focusing of the electron pattern at the opening in the electron-multiplier assembly. The electrons passing through the opening produce, by electron-multiplier action, a much larger output current.

Scanning

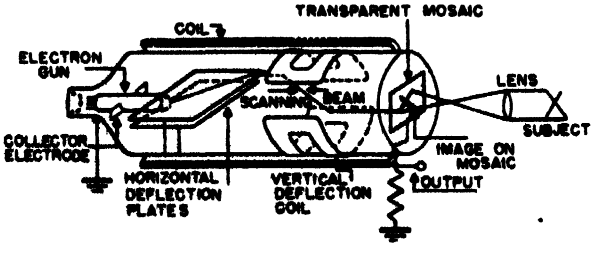
The aperture or opening in this tube is stationary. It is, therefore, necessary to move the electron pattern past the aperture in a manner that will allow electrons to be taken from the pattern. This occurs in proper sequence of picture elements and scanning lines.

Vertical- and horizontal-deflection coils, external to the tube, deflect the electron image past the opening, according to the scanning action desired. The electron-multiplier output current represents the desired camera signal and possesses all the information gathered from the video elements of the electron image.

The resolution of this type tube is excellent, but its comparatively low sensitivity has resulted in its gradual abandonment, except for film pickup where sufficient light intensity is available.

Orthicon

This is another type, differing from the image orthicon discussed earlier in this chapter, which is virtually obsolete. Its operation can be understood by reference to Fig. 21-18, and the following explanation.



Courtesy Video Handbook

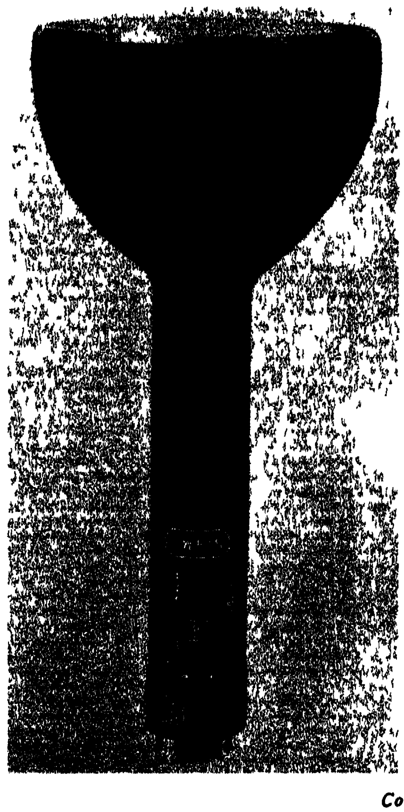
Fig. 21-18.—Early type of image orthicon.

Basically, the orthicon is a storage type. It is similar, in some respects, to the iconoscope. The optical image is focused on a transparent plate which has a photosensitive coating (mosaic) on one side. The electron beam, within the tube, is made to scan the mosaic. Light rays striking the mosaic cause changes in mosaic surface potential by photoelectric action. The beam striking a particular point returns the point to an equilibrium condition. The number of the electrons and the electron output current of the tube bear a definite relationship to the incident light energy. The video signal is thus derived.

The physical arrangement of the tube is convenient. No keystoning effect occurs, since the optical signal and the electron scanning beam strike the mosaic along the main axis. The tube is compact and small lenses may be used. Its sensitivity is 10 to 20 times that of the iconoscope but its resolution is poorer and noise level higher.

Flying-Spot Cathode-Ray Tube[5]

This type of tube is illustrated in Fig. 21-19. A block diagram of the associated circuit arrangement is shown



Courtesy RCA

Fig. 21-19.—Flying-spot cathode-ray tube.

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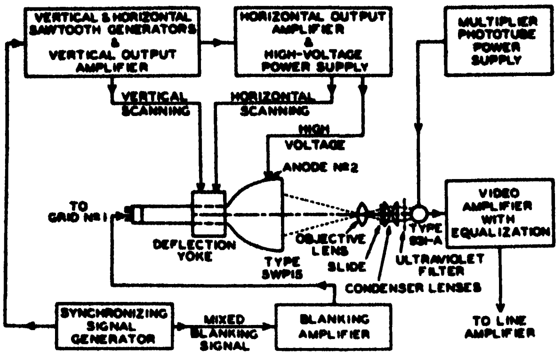
[5] This material on the flying-spot cathode-ray tube consists of excerpts from RCA Tube Data Sheet, 5WP15, 1948.


VERTICAL & HORIZONTAL SAWTOOTH GENERATORS & VERTICAL OUTPUT AMPLIFIER
HORIZONTAL OUTPUT AMPLIFIER & HIGH-VOLTAGE POWER SUPPLY
MULTIPLIER PHOTOTUBE POWER SUPPLY
VERTICAL SCANNING
HORIZONTAL SCANNING
HIGH VOLTAGE
ANODE No2
TO GRID No1
DEFLECTION YOKE
TYPE 5WP15
OBJECTIVE LENS
SLIDE
CONDENSER LENSES
ULTRAVIOLET FILTER
TYPE 931-A
VIDEO AMPLIFIER WITH EQUALIZATION
TO LINE AMPLIFIER
SYNCHRONIZING SIGNAL GENERATOR
MIXED BLANKING SIGNAL
BLANKING AMPLIFIER

Courtesy RCA

Fig. 21-20.—Block diagram of the flying-spot cathode-ray tube and its associated circuit.

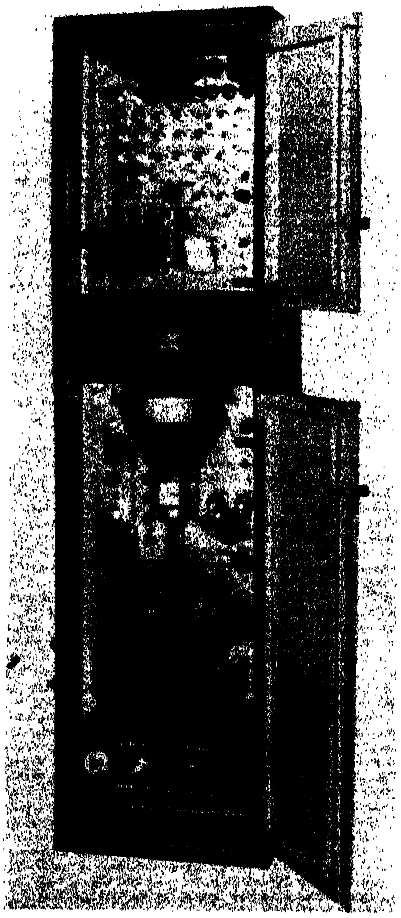
in Fig. 21-20. The cathode-ray tube is used as a source of light. This light is directed, through an objective lens, to a transparent slide. The light then passes, through condenser lenses and an ultraviolet filter, to a phototube which converts the light signals into electrical impulses. The video amplifier than builds up the signals and feeds them into the line, as shown in Fig. 21-20.

Vertical and horizontal magnetic-deflection circuits cause a movement of the beam and of the "flying spot" on the short-persistence cathode-ray-tube screen. Electrostatic focusing is used. A photograph of a scanner system similar to the one described, made by Du Mont, is shown in Fig. 21-21.

The light output of the tube is relatively low in value and is held reasonably constant during scanning. The density of the slide or film material affects the transmission of the light. Shading is thus possible. Optical limitations are disadvantages but the high scanning speed required for high-definition pictures is easily obtained.

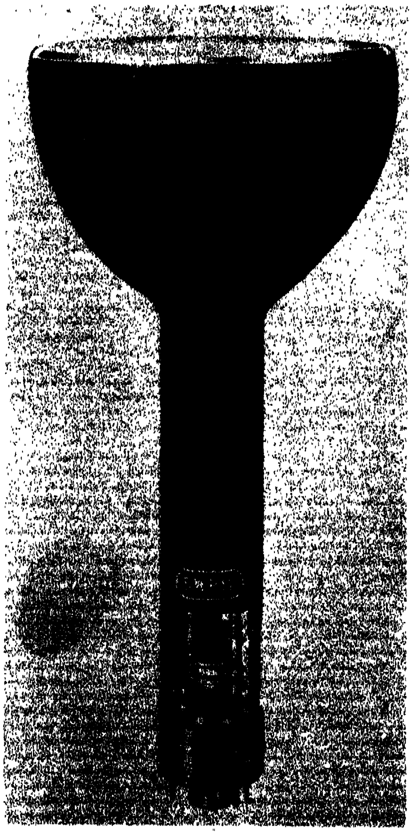
Transcriber Kinescope[6]

A tube of this kind is shown in Fig. 21-22. The fluorescent screen emits highly actinic blue radiation which is effective for photography. The tube is intended pri-



Courtesy Du Mont Labs.

Fig. 21-21.—Du Mont scanner, type TA-150-A.



Courtesy RCA

Fig. 21-22.—Transcriber kinescope.

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[6] The material on the transcriber kinescope is abstracted from RCA Tube Data Sheet, 5WP11, 1948.

marily for use in equipment which photographically transcribes television broadcasts on motion picture film. The video signal fed into the tube results in an image that can be photographed efficiently.

The persistence of the screen radiation is sufficiently short to prevent "carry-over" from one frame to the next. The metal backing of the screen effectively doubles the radiant energy output in comparison with that obtained when no metal backing is used.

Because of its large, highly actinic output, the tube can be operated at a low value of anode No. 2 current to obtain sharpness as well as ample radiation to insure adequate exposure with a medium-speed lens and with ordinary (blue-sensitive) orthochromatic or panchromatic film.

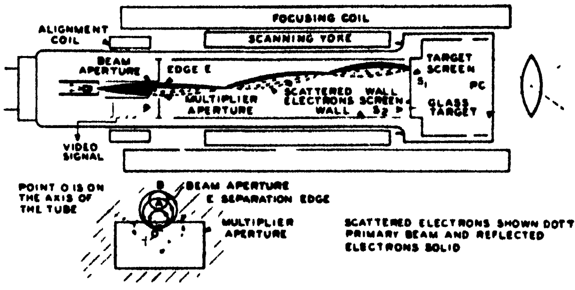
High-quality optical glass is used for the flat tube face. It will not limit the performance of a high-quality objective lens needed in the photographic equipment to provide maximum resolution.

In the design, magnetic deflection and electrostatic focus are utilized to permit obtaining essentially uniform focus over the useful screen area. Other design features include (1) an external conductive coating on the neck which, when grounded, prevents corona effects between yoke and neck, (2) a built-in capacitance between interior and exterior neck coatings to serve as a filter capacitor for the high-voltage power-supply unit, (3) an external insulating coating on the bulb cone to minimize sparking over the glass bulb under conditions of high humidity and (4) the duodecal 7-pin base designed for high-voltage service. Certain precautions in the operation of the tube must be observed to minimize the possibility of failure due to humidity, dust, and corona.

Isocon[7]

The isocon is similar, in some respects, to the image orthicon. At the moment it is purely an experimental tube, but it may assume commercial importance in the near future. A sketch of the tube appears in Fig. 21-23, If it also has an image section, the tube is called an *image isocon.*

The isocon method of signal generation is based upon the fact that the low-velocity beam impinging on the target produces scattered electrons whose number is greatest in the positive (or bright) parts of the target. By means of electron-optical techniques, the scattered electrons may be separated from the bulk of the return beam and these alone admitted to the electron multpilier.



Courtesy RCA Rev.

Fig. 21-23.—Sketch of the image isocon.

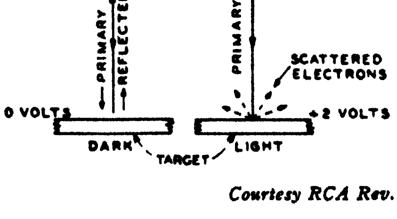
Electron Scattering

An electron beam impinging on a metal plate or on an insulator, with only a volt or two of bombarding energy, is not completely absorbed. The secondary-emission ratio for metals, instead of being zero, is of the order of a few tenths and for insulators may be as high as eight or nine tenths. True secondary emission does not occur unless the bombarding energy exceeds some critical voltage characteristic of the material, of the order of 5 to 10 volts. For voltages below this value, the electrons escaping from the surface are primary electrons which are reflected without energy loss.

Electron reflection occurring from a glass surface need not be similar to reflection from a mirror, which is often called *specular* reflection. An electron striking the surface normally will have an angle of reflection anywhere from zero up to 90°. For 1-volt incident velocity, all angles of reflection are equally probable for electrons impinging upon a fluorescent screen.

The term "scattered electrons" refers to nonspecularly reflected electrons (i.e., angles of reflection may differ from angles of incidence), obtained when a low-velocity beam impinges on a surface. A simplified picture illustrates this point in Fig. 21-24. The remainder of the returning electrons are termed "reflected electrons." This term includes those electrons which,

Fig. 21-24.—Simplified picture of electron scattering at the target of the image isocon.



Courtesy RCA Rev.

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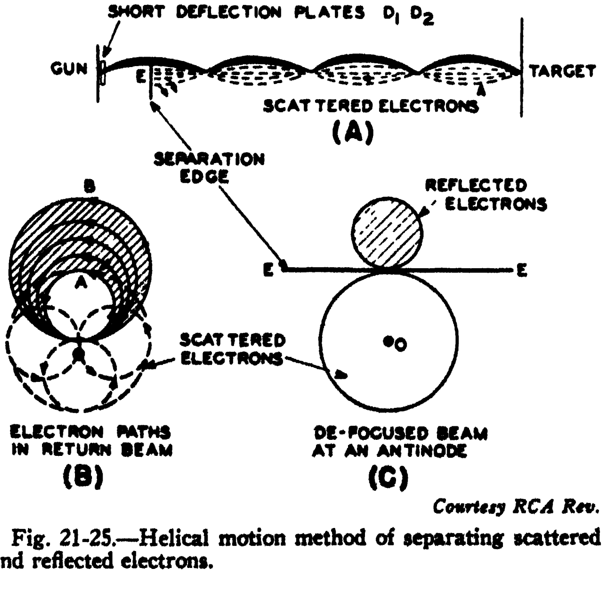
[7] Weimer, Paul K., "The image isocon — an experimental television pickup tube based on the scattering of low-velocity electrons," *RCA Rev.*, vol. 10, pp. 366-386, September 1949; the material on the isocon, appearing in this section, consists of excerpts from this article.

having insufficient energy to reach the target surface, are specularly reflected in the space in front of the target. In fact, the latter group may compose the bulk of the reflected electrons if the energy spread of the incident beam is as much as 1 volt or greater.

Separating Scattered and Reflected Electrons

An efficient method for separating the scattered and reflected electrons is illustrated in Fig. 21-25. This method consists of deliberately introducing additional helical motion into the primary electron beam so that the scattered electrons of the return beam are all found on one side of the reflected electrons. In Fig. 21-25 the excess helical motion is produced by a set of short deflection plates mounted close to the gun. The plates give a fixed transverse velocity to all the primary electrons which, when superimposed on their own random velocities, results in the helical paths whose cross section is shown in somewhat exaggerated form in Fig. 21-25. Electrons now approach the target obliquely along helices whose diameters may range from *OA* to *OB*. The transverse energy of electrons in the primary beam may range from 1 volt up to 4 or more. (The addition of a fixed transverse velocity to all the electrons in the beam increases the existing energy spread of the beam several fold.) Since this transverse energy is acquired at the expense of the longitudinal energy, the "cutoff" or black potential of the target is raised by 1 volt positive with respect to the gun cathode potential.

When a small amount of light falls on the tube, raising the target potential slightly above its cutoff value, the first group of electrons able to land on the target will be those with the least helical energy. These electrons exhibit helical motion in a diameter indicated by *OA* of Fig. 21-25B. On striking the target, they are scattered in all directions, forming many helices whose diameters may be any size from zero up to *OA*. Since all the primary electrons were focused on the point *O* at the instant of impact, the magnetic line through this point forms the central axis of the scattered beam. Fig. 21-25C shows the cross section of the return beam at an antinode. At the antinodes, all of the scattered electrons fall below the line *E-E* for the low target potential specified, while the reflected electrons are all found above this line.



Courtesy RCA Rev.

Fig. 21-25.—Helical motion method of separating scattered and reflected electrons.

Group Separation

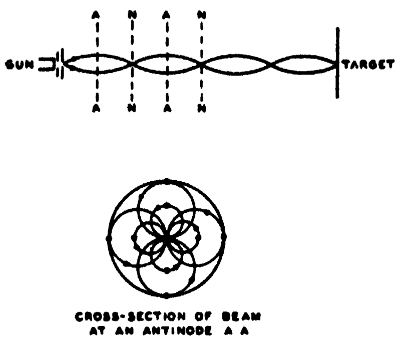
The two groups may be completely separated as shown in Fig. 21-25A by placing an electrode so that it intercepts the scattered electrons, while letting the unwanted reflected electrons pass by. Or, a better method, used in the experimental image isocon, is to admit the scattered electrons into an electron multiplier while intercepting the reflected electrons on the edge of the multiplier aperture.

For higher target potentials, corresponding to the brightest parts of the picture, a small fraction of the scattered electrons fall above the line *E-E* and are lost. This is the consequence of the wider scattering produced by the increased impact energy. It is advisable to operate so that the separation edge *E-E* passes approximately through *O*, the center of the scattered beam. Under these conditions, at all light levels, half of the scattered electrons are utilized and the linearity of response is preserved.

Operation

The helical motion of the primary beam necessary for separation of scattered electrons is obtained in the case of the experimental image isocon shown in Fig. 21-23 by positioning the aperture labeled "beam aperture" slightly off the axis of the tube. This aperture is approximately forty mils in diameter and is located at an antinode in the beam. The beam emerging from the defining aperture is made up of electrons having all directions of helical motion (Fig. 21-26). The beam aperture at the first antinode selects out a group of electrons to form a primary beam having the proper helical motion for separation of the scattered electrons.

A second aperture labeled "multiplier aperture" is also in the plane of the antinode and located relative to the beam aperture so that it passes only the scattered electrons in the return beam. The first stage of the multiplier is mounted immediately behind the multiplier



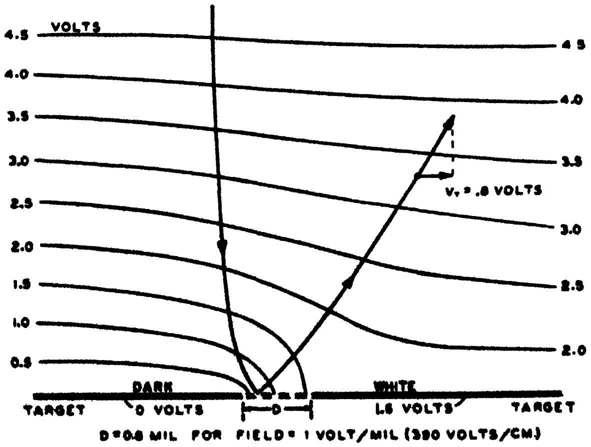
Courtesy RCA Rev

Fig. 21-26.—Electron paths in a uniform magnetic field.

aperture. The upper edge of the multiplier aperture serves as the separation edge *E-E*. The reflected electrons are intercepted on this edge of the multiplier aperture and do not enter the multiplier. The output signal from the multiplier is modulated with the opposite polarity to that of the image orthicon, having maximum current in the light instead of in the dark.

In addition to the special apertures for the gun and multiplier, the wall screen *S2* forms an essential part of the tube. This screen is of fine mesh, and, being mounted at an out-of-focus position, is not resolved by the beam. The strong uniform electric field produced in front of the target by this screen is necessary to minimize the residual scan of the return beam, and to reduce beam bending in the vicinity of the target. The latter effect may give rise to spurious edges in the transmitted picture.

The need for restricting the scan of the return beam is essential to successful separation of the scattered electrons. If the return beam moves appreciably, relative to the separation edge *E* some of the reflected electrons may enter the multiplier and give spurious shading. Inasmuch as the deflection fields serve only to bend the magnetic lines of the focusing field, it might be expected that the return beam would exactly retrace the path of the incident beam.



Courtesy RCA Rev.

Fig. 21-27.—Electron path in the neighborhood of the target (low velocity scanning). The path length shown is only a few thousands of an inch in length and the effect of the magnetic field over this distance is negligible.

The need for a strong electric field in front of the target of an image isocon is a consequence of the method of separating the scattered electrons. As indicated in Fig. 21-27, an electron approaching the target at a boundary between a white and dark area will be deflected toward the more positive area.

The transverse impulse at the target causes the electrons to follow a different diameter helix on their return path and results in their displacement relative to the separation edge. For certain orientations of the black and white boundary at the target, the reflected electrons may be deflected into the multiplier aperture, giving a spurious bright edge on the boundary of the transmitted picture.

To minimize this effect, the electric field in front of the target should be at least 500 volts per centimeter. Even with this field some beam deflection will occur and the transmitted picture will appear slightly differentiated if the selection edge *E* is located as shown in Fig. 21-25. This procedure eliminates the spurious edge effect but reduces the output signal by one half.

Operating Adjustments

Theoretically, the scattered-electron tube shown in Fig. 21-23 should not require any special adjustments when first inserted in the camera coil and power is applied. The multiplier aperture is so located relative to the beam aperture that only scattered electrons should enter the multiplier, giving immediately the proper polarity signal. Unfortunately, the electron-optical conditions are sufficiently critical that a slight misadjustment may produce either no picture at all or a picture of wrong polarity. The most likely causes for a poor initial picture are (1) bad mechanical alignment of the tube or coil (2) improper voltages or fields to give an antinode in the plane of the separation aperture, or (3) too much residual scan in the return beam.

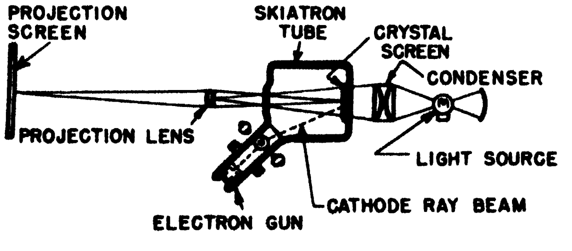
In experimental tubes, misalignment may be the principal offender. An alignment coil (Fig. 21-23) over the gun, giving a small transverse field of variable direction and magnitude, is the most satisfactory means of correcting for existing misalignments.

The alignment should be adjusted while watching the picture. The principal effect of the alignment is to center the beam properly over the beam aperture, rather than to give it helical motion. A widely diverging beam from the defining aperture is necessary, to be sure that when the beam is properly positioned over the aperture to select electrons of correct helical motion there will be some electrons having that motion.

The present method of separation of scattered electrons is based on having an antinode in the beam in the plane of the separation edge. Thus, a particular ratio of gun voltage to magnetic field strength is required to match the tube dimensions. Nonuniform shading in the video signal of the image isocon electrons can nearly always be traced to too much residual scan in the return beam. Although the decelerating screen *S2* in Fig. 21-23 greatly reduces this scan, it may still be excessive unless proper operating precautions are taken. The beam on its initial and return passage along the curved magnetic lines of the deflection coil will acquire additional helical motion which may cause the return beam to scan over the edge of the multiplier aperture. This scan may be eliminated by adjusting the distance of the deflection coil from the target so that the helical motion induced in the incident beam by the deflection coil will be cancelled by the effect of the deflection coil on the return beam.

The Skiatron

The skiatron is a television picture tube developed especially for television projection systems. In this application it is far more efficient than the conventional fluorescent-screen tubes. This is largely due to the fact that a diminution rather than an intensification of the light output of the tube is utilized in forming the image. Thus much better definition is made possible.



Courtesy Scophony Corp. of Amer.

Fig. 21-28.—Diagram of the skiatron and its associated projection system.

The tube together with the rest of the projection system is shown in Fig. 21-28. The "dark" image that forms on the crystal screen results from the following action. The screen, which is a mosaic of crystals (usually potassium chloride), darkens at the area struck by the electron beam. The intensity-modulated electron scanning beam will then produce a "negative" of the original image, provided the phase of the signal is maintained properly.

This negative image is projected as it is formed onto the projection screen where it appears in its final "positive" form. It will be noted that unless it is projected onto the screen, the "dark" image is unsuited for viewing purposes, unlike conventional projection systems.

The mechanics of the image-forming process on the crystal screen is not understood too well. It is believed that when the electron beam strikes a crystal element of the mosaic, which was originally fairly transparent to the light from the source, the crystal becomes polarized in such a way as to become opaque. The degree of opacity is proportional to the intensity of the electron beam.

Visual Indicator Tubes

A visual indicator tube, which is sometimes called a tuning indicator, is essentially a special type of cathode-ray tube. Although it is frequently used on test equipment as an indicator, its most common use is as a tuning indicator that shows when a receiver is tuned in properly on a radio station.

As is shown in Fig. 21-29, a visual indicator tube consists of a triode *P*, *G*, and *K* and a modified cathode-ray tube, *K*, *V*, *KS*, and *T* in the same glass envelope. The cathode extends beyond the triode plate section and also serves as a source of electrons for the cathode-ray section. A cathode shield *KS* is positioned in such a manner as to prevent the red-hot cathode *K* from being seen from the top of the tube. A thin metal ray-

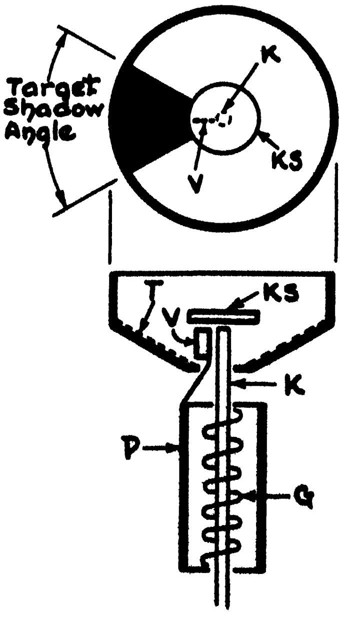


Fig. 21-29.—The arrangements of the elements of a visual indicator tube.

control vane *V*, which functions to control the opening of the shadow angle, is located between one side of the cathode *K* and the fluorescent target *T*, and is electrically connected to the triode plate *P*. The fluorescent target is inclined at an angle to the axis of the cathode.

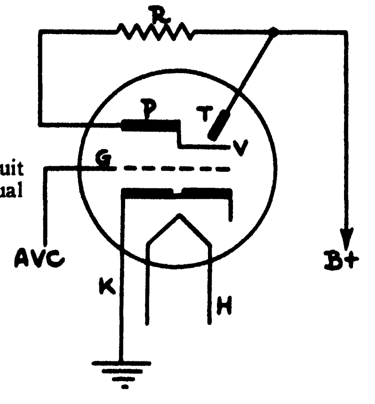


Fig. 21-30.—The circuit connections for the visual indicator tube.

The target *T* is connected directly to the power-supply voltage, as shown in Fig. 21-30, and a resistance *R* is connected between the target and the triode plate *P*. The plate and the ray-control vane *V* are consequently always at a voltage that is less than that of the fluorescent target by an amount equal to the voltage drop across the plate-target resistor *R*. This voltage drop depends on the magnitude of the plate current, which, in turn, is dependent upon the value of the grid voltage.

Suppose that a sufficiently high negative voltage is applied to the grid *G* to drive the plate current beyond its cutoff value. Since no plate current flows, there is no voltage drop across the plate-target resistor. The triode plate and ray-control vane are, therefore, at the same potential as the target. Now electrons are being emitted by the hot cathode in all directions and since the fluorescent target is positive with respect to the cathode, the target attracts these electrons. As a result of this electron bombardment, the entire fluorescent target becomes illuminated with the X-ray-control vane. The appearance of the target when the triode grid is highly negative is shown in Fig. 21-31.

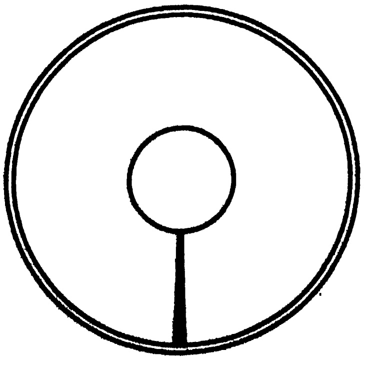


Fig. 21-31.—When the triode grid of the indicator tube is highly negative, the shadow on the target is narrow.

Now assume that the grid is no longer negative, but instead has zero volts applied. Plate current will flow and a large voltage drop appears across the plate-target resistor. The positive voltage of the triode plate and, therefore, of the ray-control vane is now considerably lower than the positive voltage of the target by an amount equal to the voltage drop across the plate-target resistor. Although the ray-control vane is positive with respect to the cathode, this vane is now highly negative with respect to the target. The electrostatic field between the ray-control vane and the target is, therefore, of such character as to result in a repulsion of the electrons emitted by the cathode in a direction away from the ray-control vane. Consequently, no electrons will strike the fluorescent target in the vicinity of the ray-control vane, so that this portion of the target will be dark. The appearance of the target for zero volts on the triode grid is shown in Fig. 21-32.

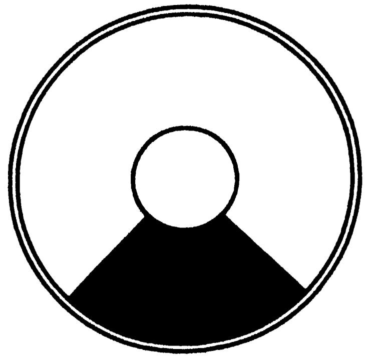
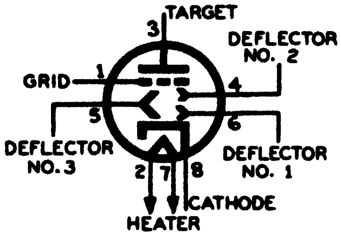


Fig. 21-32.—As the voltage on this grid approaches zero, the shadow widens as shown.

The negative control voltage applied to the triode grid (or the lack of it for the zero-voltage condition) may be obtained from a circuit in a radio receiver, this circuit being known as the automatic-volume-control or avc circuit. In a circuit of this type, a highly negative voltage is developed when a radio station is tuned in accurately, and if this negative voltage is applied to the grid of a visual indicator tube, it will be seen that the shaded sector will reduce to a narrow line as indicated in Fig. 21-31. On the other hand, if the receiver is not accurately tuned to the radio station, then very little (or no) negative voltage will be developed by the avc circuit and the angular sector of the visual indicator tube will appear as shown in Fig. 21-32.

An F-M Tuning Indicator

Late in 1946 the General Electric Company released a new tuning indicator which worked on cathode-ray principles and was suited ideally for f-m receivers, although it can also be used on a-m receivers. This type tube is known as a 6AL7-GT, a schematic of which appears in Fig. 21-33.



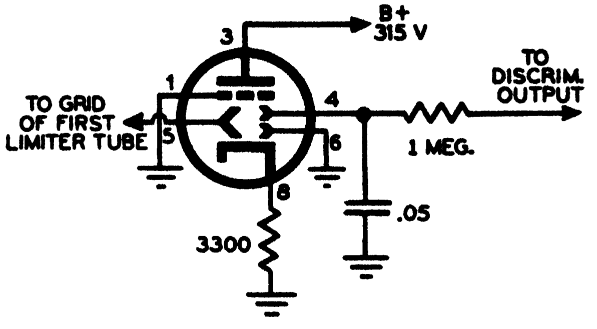
Courtesy GE

Fig. 21-33.—The schematic of the 6AL7-GT visual indicator tube.

This tube can make use of both the discriminator output voltage and the limiter grid voltage. It can be used satisfactorily with just the discriminator voltage alone, but it then is not able to distinguish whether the receiver is on tune or off channel. For, when the discriminator output voltage is used alone, the pattern for on tune or off channel is the same because the discriminator output voltage is zero in both cases.

The tube is made sensitive enough to respond to a voltage difference between ±0.2 volt with respect to ground; this is within 2 kc of the discriminator tuning for distortionless signals. This discriminator output is connected to one side of a center-divided deflector (deflector *No. 2* in Fig. 21-33). A space-charge grid is used to increase the sensitivity of the tube. The deflector is so divided to form a method whereby the deflection of one half of the pattern can be compared with the other half. The other half of the divided deflector is usually grounded (deflector *No. 1* in Fig. 21-33), providing a reference pattern with which the pattern due to deflector *No. 2* may be compared.

Since one side of the discriminator load is usually grounded, the changing discriminator output voltage appears across both these deflector electrodes. On one side of the cathode is placed this divided deflector, and on the other side another deflector is placed to form a fixed boundary for one side of the pattern (deflector *No. 3* in Fig. 21-33). The first limiter grid (if more than one limiter is used) is connected to this electrode, and the limiter voltage, therefore, appears on this deflector and is used as a voltage which determines one boundary of the target pattern. This deflector enables us to distinguish, by different patterns, when the receiver is on tune or off channel. A circuit diagram of how the tube usually is wired for operation in conjunction with a limiter and discriminator is shown in Fig. 21-34.



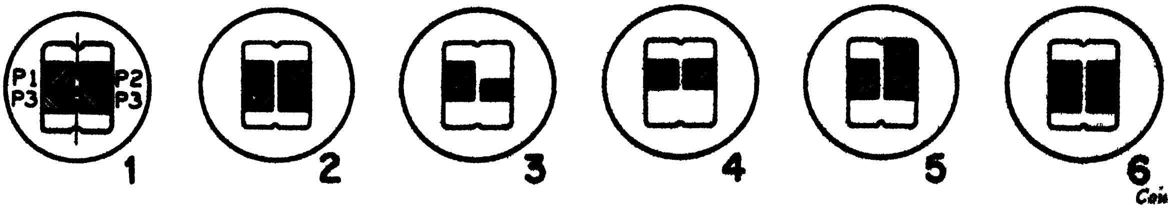
Courtesy GE

Fig. 21-34.—The 6AL7-GT connected in a f-m receiver employing a limiter.

In Fig. 21-35 are illustrated the target patterns under different conditions of discriminator output. In pattern *1* four squares are shown where *P1, P2,* and *P3* are produced and controlled by deflection electrodes *No. 1, 2,* and *3* respectively; there are two *P3* squares. Patterns *2* and *6* are for minus and plus off-channel conditions, respectively. These patterns are identical because there is zero voltage output from the discriminator and no limiter voltage. The size of squares *P3* is controlled by the amount of limiter voltage; the higher this negative voltage the smaller the squares, and the lower the negative voltage, the larger the squares. Consequently, in patterns *2* and *6* the bottom half is of maximum depth due to the absence of limiter voltage.

In pattern *4* the receiver is on tune, which means a maximum negative limiter voltage and a zero discriminator output voltage. Thus the pattern is much smaller than those off-channel patterns of *2* and *6*. The decrease in pattern *4* is seen to be in the bottom half due to the limiter voltage.

In patterns *3* and *5* where the discriminator is off tune, limiter voltage, although perhaps not a negative maximum, is still present, limiting the bottom half of



Courtesy GE

Fig. 21-35.—Various target patterns under different conditions of discriminator output.

the patterns to a smaller size as compared with the patterns of *2* and *6*. In pattern *3* the discriminator is off tune and presents a negative signal to the *No. 2* deflector grid, making square *P2* reduced in size. In pattern *5* the discriminator is also off tune and presents a positive signal to the *No. 2* deflector grid, making square *P2* increased. You will note that the deflection due to the positive off-tune signal is greater than the deflection for the negative off-tune signal under the same amount of positive and negative discriminator voltages. This is a result of the use of space-charge operation of the deflection system (due to the space-charge grid). Deflection in the positive region would be much greater if a cathode bias resistor (3,300 ohms as seen in Fig. 21-34) were not used. This resistor places a positive voltage on the cathode with respect to the deflectors and the space-charge grid, so that with a positive discriminator signal the deflectors do not draw appreciable current.

The negative bias on the space-charge grid reduces the brightness of the pattern and increases the deflection sensitivity.

CHAPTER 22

COMMERCIAL OSCILLOSCOPES AND RELATED EQUIPMENT

All types of oscilloscopes and similar equipment employing cathode-ray tubes will be considered in this chapter. The important specifications, such as horizontal- and vertical-amplifier responses, sweep speeds, deflection factors, and power rating, are given for each instrument. In addition, there is a description of each instrument, in greater or lesser detail, depending on the existence of any unique features in the particular instrument. Most of the conventional, general-purpose oscilloscopes have little, if any, description in addition to the specifications because the information about such circuits has been discussed in previous chapters of the book.

The special-purpose instruments, such as synchroscopes, hydrauloscopes, engine analyzers, signal-tracing, and other servicing oscilloscopes, etc., are also included in this chapter and their circuits are discussed commensurate with their complexity and departure from conventional forms.

The schematic diagram is given for each piece of equipment, as well as its tube complement. Where the tube data is included in the schematic, it is omitted from the specifications. The instruments are listed alphabetically by manufacturer and, under each manufacturer, by model number and letter.

In these analyses, the attempt has been made to be as concise as possible from considerations of space and convenience and yet present sufficient information to identify the instrument and its potentialities.

AEROQUIP MODEL 10,000

FREQUENCY RESPONSE

Vertical Amplifier Flat, d.c. to 10,000 cps, 1 db down at 20,000 cps

Sweep Circuit 15, 3, 1, 0.3, 0.1, 0.03, 0.01, 0.003, 0.001 seconds

LINE RATING 105-125 volts, 50-1,800 cps

TUBE COMPLEMENT

Type	*Function*
12AH7-GT (V1A)	Y-Amplifier Input
12AH7-GT (V1B)	X-Amplifier
12SL7-GT (V2A)	2nd Y-Amplifier
12SL7-GT (V2B)	Y-Amplifier Output
2050 (V3A)	Single-Trace Control Tube
2050 (V3B)	X-Axis Sweep Generator
OD3 (V4)	Voltage Regulator
OC3 (V5)	Voltage Regulator
OA3 (V6)	Low-Voltage Regulator
12SJ7 (V7A)	Timing-Wave Generator
12SJ7 (V7B)	Low-Voltage Regulator
6AS7 (V8)	Low-Voltage Regulator
5U4-G (V9A)	Medium-Voltage Rectifier
5U4-G (V9B)	Low-Voltage Rectifier
2X2/879 (V10)	High-Voltage Rectifier
6V6-GT (V11)	High-Voltage Regulator
5BP1 or 5LP7 (V12)	Cathode-Ray Tube

The schematic circuit diagram of the Aeroquip Hydrauliscope is shown in Fig. 22-1. This is not an ordinary oscilloscope such as might be used in an electronic laboratory or in radio servicing, but it was designed specifically for hydraulic pressure measurements. It provides visual presentation of the variation in pressure in a hydraulic system with respect to time or motion. It will respond faithfully to static or transient phenomena up to a frequency of 20,000 cps. The input circuit may be modified, externally or internally, to accept the output of practically all the pressure pickup elements which convert changes in some variable to changes in electrical potential.

Pressure Pickup Element

This element includes two arms of the Wheatstone bridge circuit. One of these arms consists of a winding of one mil diameter wire-wound under tension on an especially prepared metal tube. Pressure inside the tube causes an increase in its diameter, thereby stretching the wire and increasing its resistance. Different operating pressure ranges are obtained by varying the wall thickness of the tube. The second arm of the bridge is a winding identical to the first, but not subject to pressure changes. This winding is used for temperature compensation.

Time Bases

The horizontal or X-axis time-base circuit consists of a 2050 thyratron *V3B* and an *R-C* charging circuit. Nine fixed time bases are provided, and are selected by a switch. The rates range from 15 seconds to 1 millisecond per sweep of 3.5 inches. This time-base generator is connected directly to the X-TIME amplifier which uses a 12AH7-GT tube *V1B* feeding the horizontal-deflection plates of the cathode-ray tube. The starting position of the trace is controlled by the X-POSITION control *R19*.

X-Selector

The generator *V3B* may be triggered in different ways, as selected by the X-SELECTOR switch *S4*. In the CONTINUOUS position, the grid bias of *V3B* is adjusted by *R31* to the point where the thyratron acts as a relaxation oscillator. These oscillations may be synchronized, if desired, by the application of a signal between terminals *A* and *B* of the EXTERNAL receptacle *J2*.

In the INTERNAL REPEAT position, sufficient bias is supplied to *V3B* so that it will not ionize until it receives a signal from the Y amplifier through *C3A*. This signal must be positive and must correspond to an upward deflection greater than one-quarter inch.

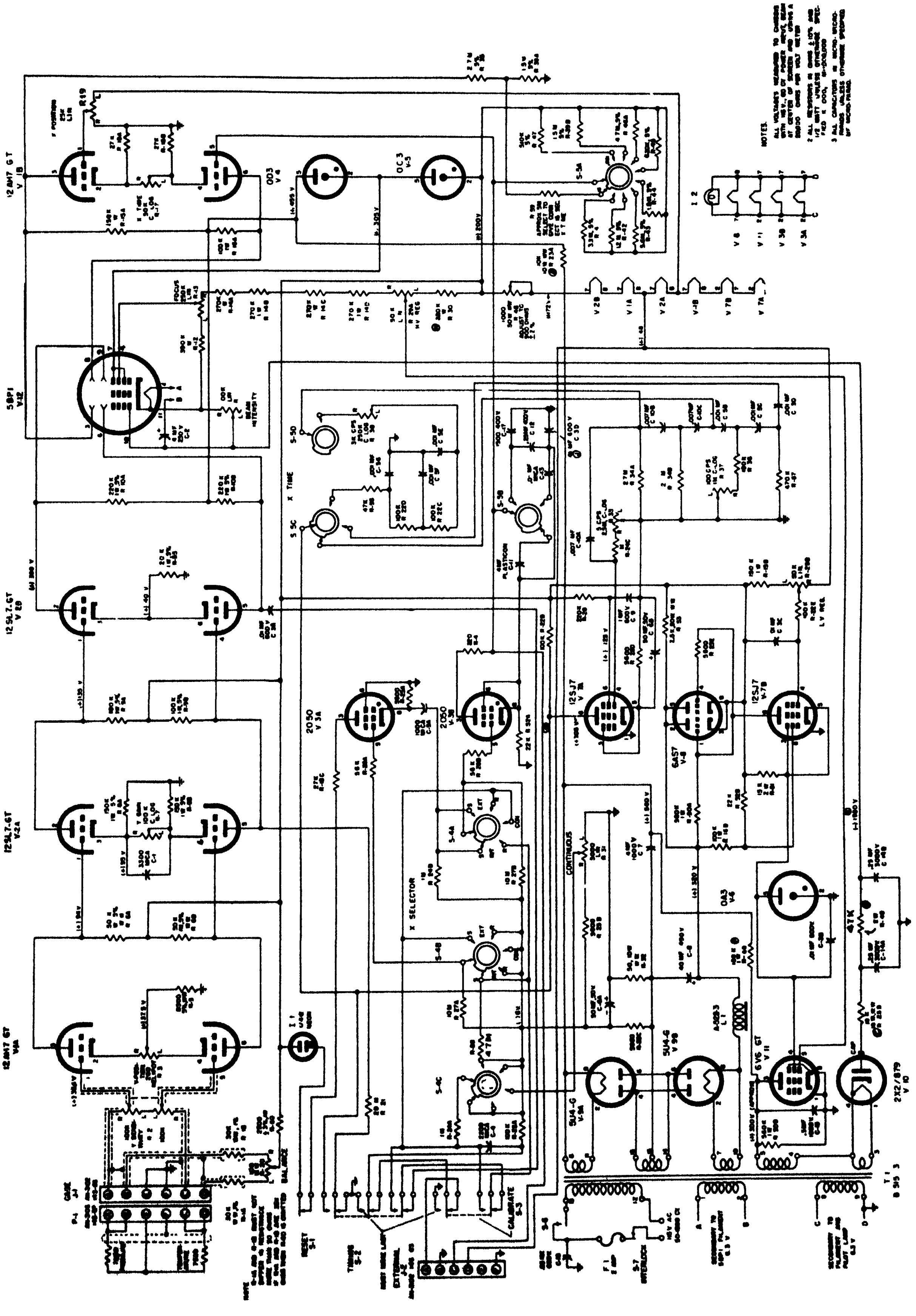
In the EXTERNAL REPEAT position, *V3B* may be triggered by simply shorting terminals *A* and *B* of the EXTERNAL receptacle *J2*. This is usually done by means of a suitable contactor operated from some element in the hydraulic system. Each time the contactor closes, *V3B* will receive a trigger pulse. The EXTERNAL SINGLE position of the X-SELECTOR switch connects the second thyratron *V3A* between the EXTERNAL receptacle *J2* and the grid of *V3B*.

When the contactor closes, *V3A* conducts and sends a signal to *V3B*. Since there is no capacitor in the anode circuit of *V3A*, this thyratron will remain conductive regardless of the opening and closing of the external contactor until the anode circuit is broken by pressing the reset button *S1*.

Since *V3B* can receive a signal only at the instant *V3A* begins to carry current, it is apparent that only one trace will be drawn unless the reset button is operated. In the INTERNAL-SINGLE position, operation is the same as the EXTERNAL-SINGLE position except that the trace is initiated by a positive signal from the Y-axis amplifier. A sweep of the beam may also be initiated in the EXTERNAL position by a pulse of positive voltage (six volts minimum) applied to terminal *A*.

Timing-Wave Oscillator

In order to provide accurate indication of time intervals on the X-axis, a stable resistance-capacitance coupled oscillator is included. This circuit uses a type 12SJ7 tube *V7A* and is arranged so the frequency of oscillation changes with the X-TIME switch *S5*. On the 15-, 3-, and 1-second positions, the timing wave has a frequency of 5 cps. In the next three positions, 0.3, 0.1, and 0.03, the frequency of the timing-wave oscillator is 100 cps, and in the remaining three positions the frequency is 3,000 cps. The timing wave is fed to the Y-amplifier by means of the TIMING switch *S2*. When this switch is pressed, the cathode-ray beam moves down approximately one-quarter inch so that the timing wave, as it appears on the viewing screen, will be below the normal zero axis. The timing wave is intended primarily for use in connection with the camera attachment since no provision is made for synchronization with the continuous sweep.



Courtesy Aeroquip Corp.

Fig. 22-1.—Schematic of Aeroquip Model 10,000

BETA ELECTRONICS MODEL 701

Frequency Response
 Vertical Amplifier 10 cps to 50 kc, ±10%
 Sweep Circuit 25 cps to 2 kc

Deflection Factors
 Vertical Amplifier 0.06 volts rms/inch
 Horizontal Amplifier 0.065 volts rms/inch

Line Rating 110-120 volts, 50-60 cps

Tube Complement

Type	*Function*
12AU7 (V1)	Vertical Amplifier
6B6G (V2)	Vertical-Deflection Amplifier
12AU7 (V3)	Horizontal Amplifier
6B6G (V4)	Horizontal-Deflection Amplifier
6C8 (V5)	Horizontal Sweep Oscillator
5U4 (V6)	High-Voltage Rectifiers
3NP4 (V7)	Cathode-Ray Tube
1B3G (V8)	High-Voltage Rectifiers

The schematic circuit diagram for Model 701 is shown in Fig. 22-2. This instrument was designed for educational purposes, and uses conventional circuits. It has a large screen which makes it suitable for displaying waveforms to technical students, engineers, or other groups. The instrument may also be used in industrial or research applications, where cathode-ray monitoring is required, and the advantages of a large picture to reduce eye strain or to permit long-distance viewing may be required.

The cathode-ray tube is a 2½-inch type designed for projection television use. A compact folding Schmidt optical system enlarges and projects the image from the face of the tube vertically to a mirror mounted on the underside of the lid of the cabinet. This lid is raised to an angle of 45° when the oscilloscope is in use. The mirror reflects the light horizontally onto the screen plane where it is focused. The response is adequate to permit good vertical square-wave reproduction from d.c. to more than 10 kc at horizontal sweep repetition rates up to several kc. The circuit also provides for the injection of blanking or Z-axis modulation voltage into the picture.

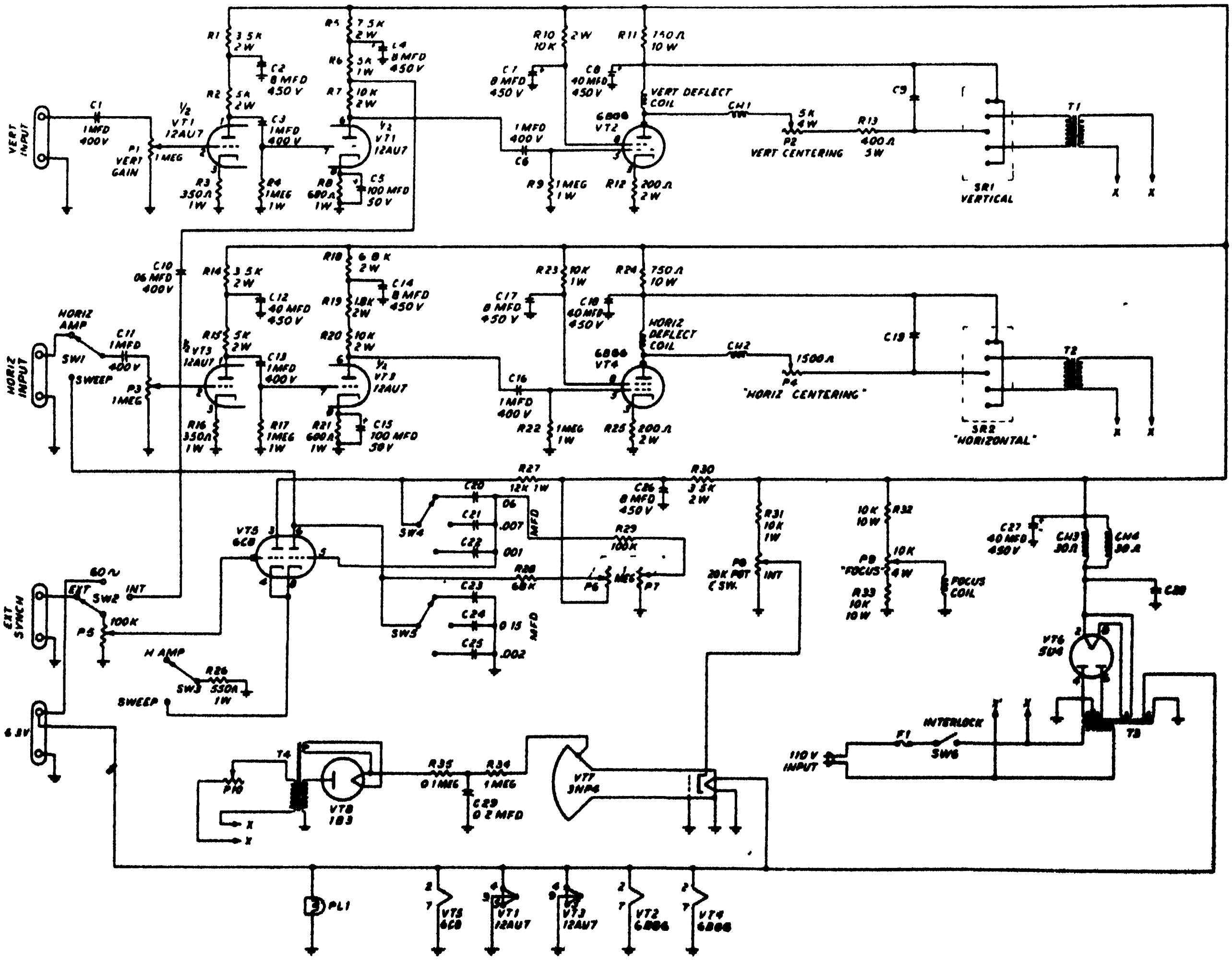


Fig. 22-2.—Schematic of Beta Model 701.

Courtesy Beta Elec. Corp.

BROWNING MODEL OL-15B

FREQUENCY RESPONSE
Vertical Amplifier 5 cps to 6 Mc, flat within 3 db
Horizontal Amplifier 5 cps to 1 Mc, flat within 1 db
Sweep Circuit 5 cps to 500 kc

DEFLECTION FACTORS
Vertical Amplifier 0.05 rms volts/inch
Vertical-Deflection Plates 100 rms volts/inch
Horizontal Amplifier 0.05 rms volts/inch
Horizontal-Deflection Plates 100 rms volts/inch

LINE RATING 115 volts, 60 cps

The schematic diagram of Model OL-15B Oscillosynchroscope is shown in Fig. 22-3. The two general classes of signals that may be studied are:

1. Signals whose duration is long compared with the time between successive signals under observation and, 2. signals whose duration is short compared to the time between successive signals.

A study of the schematic diagram, Fig. 22-3, shows several important features. The sync, sawtooth generator, and triggered sweep circuits are particularly interesting. The sync amplifier output can be fed through a switching system to the sawtooth generator, when desired, and to the triggered sweep generator; it can also be fed to the delay circuit if switched into it. The input of the triggered generator can be derived either from the delay circuit output or sync amplifier output. The triggered generator feeds into the cathode-ray-tube horizontal plate circuit, when switched in, and also into the intensifier. The intensifier goes to the first grid of the cathode-ray tube.

Input Circuit and Sawtooth Sweep

The vertical-amplifier input stage is a cathode-follower type, using *V1*; the vertical gain control is *R8*, with a series capacitor *C8* to block d.c., thereby preventing bias change of *V2* as the setting of *R8* is varied. *V2*, *V3*, and *V4* are successive amplifier stages. *V2* and *V3* use plate-circuit inductances for high-frequency compensation. *V4* is a combined voltage amplifier and phase inverter. A balanced vertical output stage of conventional design is used. Resistors *R28* and *R29* are parasitic suppressors.

The input system of the horizontal amplifier is almost identical to that of the vertical amplifier with the exception of the calibrator circuit. A 6C4 with gain control in the cathode circuit feeds a 6AG7, which, in turn, drives the push-pull 6AG7 tubes *V11* and *V12*. *V11* is a grounded-grid amplifier, in which respect it differs from its counterpart in the vertical-deflection amplifier.

The sawtooth generator is of the hard-tube variety and produces sawtooth waveforms from 5 to 500,000 per second. It is composed of *V6*, *V7*, and *V8*, with feedback action produced by *V6* and *V7*. *V8* serves as a constant current device for supplying a linear change of voltage across the sweep capacitance.

Triggered Sweep

The triggered sweep circuits in the OL-15B are built around a one-shot multivibrator consisting of a 6C4 triode and a 6SH7 triode. The gate pulse produced by this combination is amplified and inverted by *V17* first triode, and then fed to a clamper *V17* second triode, constant current pentode *V18*, and inverter *V19*, to produce a push-pull linear sweep voltage for direct application to the horizontal plates of the cathode-ray tube.

For purposes of circuit explanation, we may assume that *S8* is in position *1*, corresponding to the 0.25 microsecond per inch sweep speed condition. *C52* and *R99* are then associated with *V15* and *V16*. Since the grid return for *V16* is to the cathode, this tube will conduct readily and sufficient voltage drop will appear across *R96* to reduce the plate current in *V15* to nearly zero.

If a negative impulse should be sent through *C51*, it would be impressed equally upon the plate of *V15* and on the grid of *V16* through *C52*, resulting in a decrease of plate current in *V16*.

The decreasing space current will cause a decreasing voltage to appear across cathode resistor *R96*. Since the *V15* cathode is connected to the cathode of *V16*, the voltage change results in an effective decrease in bias on *V15* with increasing plate current. Hence, the plate voltage will drop, and this change, in turn, will be communicated to the grid of *V16*. We have a feedback condition which will result in the grid of *V16* being driven beyond the condition of plate current cutoff. *V16* remains in this condition as *C52* charges through *R99*. When the voltage on the grid of *V16* reaches a certain point, plate current begins to flow again and the voltage across *R96* increases.

Feedback action rapidly returns the circuit to its stable condition where it remains until another trigger signal arrives. Thus, a negative voltage will be produced at the plate of *V15* and a positive voltage pulse at the plate of *V16*.

C52 and *R99* are chosen to produce a gate whose duration is as long as the sweep time desired. In this case it is 1.25 microsecond (5 inches of deflection at 0.25 microsecond per inch). A portion of the positive output of *V16* is applied to the grid of *V17*, triode No. 1, which has been biased to a condition of very small plate current. As the pulse is applied, the tube conducts and a large negative pulse appears at the plate.

V17, section 2, is normally conductive so that the sweep capacitance selected by *S8C* will be charged to a voltage determined by the relative impedances of *V17*, section 2, and *V18*. When the negative voltage at the clamp tube *V17*, section 2, cuts off current in this tube, the sweep capacitance discharges through *V18* at a linear rate. This voltage is applied to the grid of *V19* through a compensated attenuator and appears as a positive sweep voltage at the plate of *V19*. Negative and positive sweep voltages are applied to the horizontal-deflection plates of the cathode-ray tube. At the same time as the sweep is applied, a positive pulse from the sweep gate is impressed on the intensity grid of the cathode-ray tube.

With SYNC SELECTOR in EXT NEG position, a negative trigger at the SYNC INPUT terminals is passed through *R88*, the SYNC GAIN control, to sync amplifier *V13*. This pulse appears as positive at the plate of *V13* and triggers the sweep gate through the sync inverter *V14A*. With SYNC SELECTOR in EXT POS position, a positive trigger at the SYNC INPUT terminals is passed through the SYNC GAIN control to the sync inverter which, in turn, triggers the sweep gate. The triggered sweep circuit may also be triggered from the video amplifier, as in the case for internal sync using the sawtooth sweep. SYNC SELECTOR will be in either the POS SIG or NEG SIG position, as the case may be.

Trigger Generator and Delay Circuit

An 884 thyratron relaxation oscillator serves as the trigger timing source in the OL-15B. A resistor *R119* of 200 ohms is in the cathode circuit of this tube *V20*. When *V20* fires, a sharp voltage pulse is formed across *R200*. This pulse is inverted by *V14B* and triggers *V21* which operates in a similar manner to that of the sweep gate multivibrator. The width of the delay multivibrator gate is adjustable by means of *R127*, the DELAY CONTROL. *R127* and *R127A* are a concentric shaft dual control with *R127* (1 meg) for coarse phasing control and *R127A* (10,000 ohms) for fine phasing control.

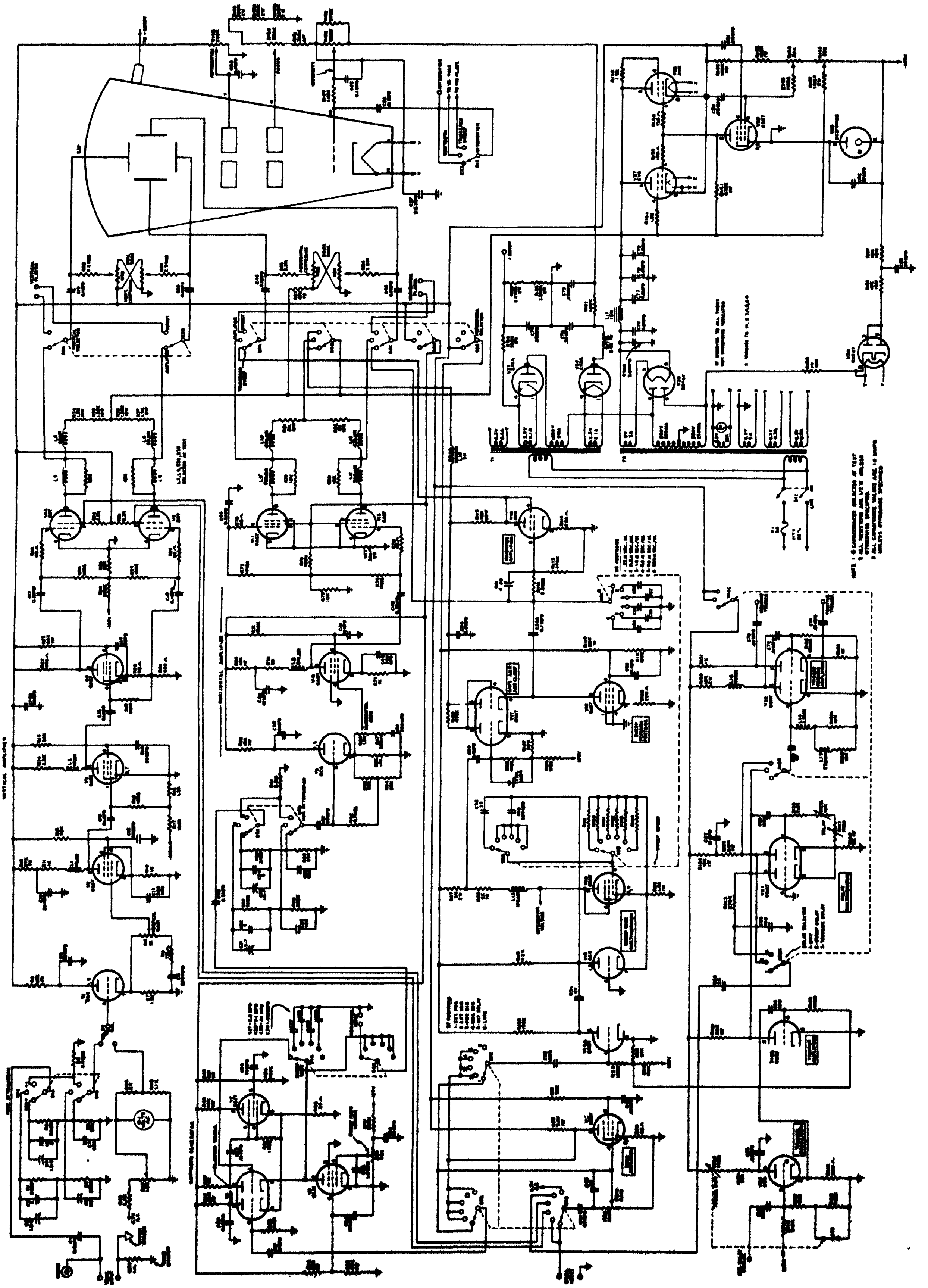
With *S10* in SWEEP DELAY position, the negative output of the gate is differentiated by the network in the grid circuit of *V22*. Since *V22*, triode No. 1, is at zero bias condition, this stage responds only to negative grid excursions so that a positive trigger voltage appears at the plate of *V22*, section 1, coincident with the leading edge of the delay gate.

At the same time, the negative trailing edge of the gate output at the plate of the second triode is applied through *C65* to trigger the sweep multivibrator through the sync amplifier *V13*.

With this arrangement, the sweep is delayed with respect to the output triggers. In TRIGGER DELAY position, the trigger is delayed with respect to the sweep. Since the width of the delay multivibrator circuit cannot be reduced to zero, it would normally be impossible to reduce the delay between the trigger and sweep to less than about 2 microseconds, so that the trigger could not be phased to occur during the two fastest sweep speeds.

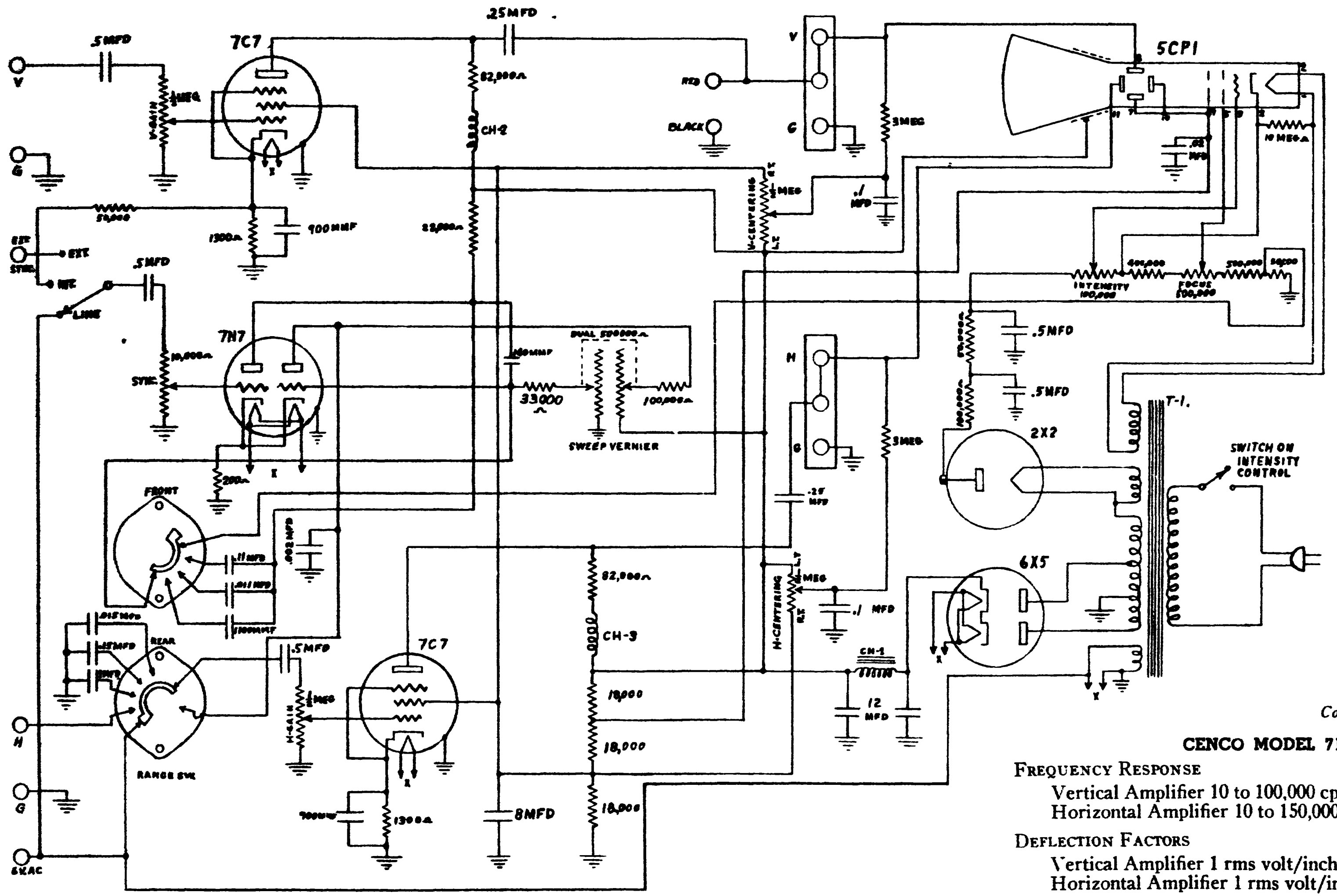
To overcome this difficulty, *R123* and *C66* are used in the circuit to delay the start of the sweep by about 2 microseconds, so that the trigger can be phased in to zero time with respect to the sweep start.

When TRIGGER RATE is turned to the extreme counterclockwise position, *S13* is opened and *V20* is heavily biased so that its running rate is very low, around 30 cps. A positive trigger of at least 30 volts applied to INT DELAY SYNC IN will trigger the thyratron, which, in turn, will operate the trigger and delay circuits in the same manner as for the free-running condition of the trigger generator.



Courtesy Browning Labs.

Fig. 22-3.—Schematic of Browning Model OL-15B.



Courtesy Central Scientific Co.

Fig. 22-4.—Schematic of Cenco Model 71552.

CENCO MODEL 71552

Frequency Response
Vertical Amplifier 10 to 100,000 cps, ±2 db
Horizontal Amplifier 10 to 150,000 cps, ±2 db

Deflection Factors
Vertical Amplifier 1 rms volt/inch
Horizontal Amplifier 1 rms volt/inch

Line Rating 110-120 volts, 50-60 cps

Tube Complement

Type	*Function*
7C7	Vertical Amplifier
7C7	Horizontal Amplifier
5CP1	Cathode-Ray Tube
7N7	Sync Amplifier and Sweep Oscillator
2X2	High-Voltage Rectifier
6X5GT	Low-Voltage Rectifier

The schematic circuit diagram for Model 71552 is shown in Fig. 22-4.

COSSOR MODEL 1035

FREQUENCY RESPONSE

Vertical Amplifier (Y1) 20 cps to 7 Mc (15% down at 7 Mc)

Vertical Amplifier (Y2) 20 cps to 100 kc (15% down at 100 kc)

Sweep Circuit 150 milliseconds to 15 microseconds

DEFLECTION FACTORS

Vertical-Deflection Plates 675 d-c volts/mm

Horizontal-Deflection Plates 800 d-c volts/mm

LINE RATING 105/115, 120/130, 200/215, 216/234, 235/255 volts, 50 to 100 cps

TUBE COMPLEMENT

Type	*Function*
43.1U (V1)	Low-Voltage Rectifier
E.F.50 (V2)	Input Tube, A1 Amplifier
61SPT (V3)	Driver
61.BT (V4)	Cathode Follower
SU2150A (V5)	High-Voltage Rectifier
63SPT (V6)	Voltage Regulator
SD6 (V7)	Clamper
63SPT (V8)	Transitron Oscillator
63SPT (V9)	Time-Base Generator
OM4 (V10)	Buffer
SD6 (V11)	Voltage Limiter
63SPT (V12)	Phase Inverter
61.BT (V13)	Cathode Follower
89D (CRT)	Cathode-Ray Tube

The schematic circuit diagram is shown in Fig. 22-5.

A1 Amplifier

The vertical amplifier uses an input tube *V2*, a driver *V3*, and a cathode-follower output *V4*, and provides negative feedback as a means of gain control. Signals to the *Y1* deflection plate normally pass through the amplifier, although direct connection to the plate is possible at the side of the instrument.

A tapped network applies negative feedback from the cathode of the output tube *V13* to the cathode of the input tube *V2* and is frequency-compensated by capacitors arranged in the plate circuit of *V3*, to avoid the inevitable phase change that would occur in *V13* if they were in the cathode circuit of that tube.

A2 Amplifier

The gain of this single-tube circuit *V4* is adjusted on the 5-volt range by the preset cathode-feedback control *R105*. The sensitivity switch, *S6*, allows five gain settings and gives control ranges from 500-0-500 volts to 5-0-5 volts for full-screen deflection; it is arranged as a frequency-compensated attenuator.

Time Base

The time-base circuit uses five tubes. A Miller integrator is coupled into a series multivibrator, and by means of range switch *S9*, nine basic scanning speeds are possible. The time base operates repetitively or may be triggered by a pulse from the A1 amplifier or from an external source. The triggering signal may be either positive or negative in sign. The polarity is selected by the sync selector switch *S16*. In the free-running condition, the time base can be synchronized by an internal pulse or, as in the case of the trigger pulse, from an external source. The selection of the polarity of the sync signal is provided by the sync selector switch *S16*.

The basic operation of the time base is as follows: Assume the bias potential on the suppressor grid of *V8* results in plate-current cutoff; *V9* is conducting (zero bias); and the capacitor selected by *S9*, say *C34*, is fully charged. At the beginning of the cycle, the suppressor of *V8* is driven positive, causing a drop in anode potential. The signal is transferred to the control grid through *C34*. This initial drop extends only for the duration of the grid base after which the anode current is limited by the negative feedback of *C34*. The *V8* anode voltage falls linearly.

The discharge current from *C34* maintains *V9* at cutoff during this period and ensures that a positive driving voltage is applied to the suppressor via *C43* (that is, there is no potential drop at the *V9* anode transferred to the *V8* suppressor through *S12* and *C43*). The scanning period continues until the linear decay in anode voltage, determined by the time constant of *C34* and *R91*, reaches the bend in the plate characteristic curve of *V8* and the discharge current from *C34* ceases.

As a result, the bias on *V9* is removed and the tube becomes conductive; its plate goes negative and supplies a cutoff voltage to the suppressor grid of *V8*. Capacitor *C34* now becomes recharged due to the current in the *V8* grid circuit and *V9* cathode circuit and is ready for the next forward stroke of operation.

The charging time of *C34* corresponds with the flyback period of the cathode-ray-tube spot. An alternative point of view is that *V8* is cut off during flyback, since *C43* is charging from negative to positive at the suppressor grid point in the circuit.

Synchronization

Synchronization is achieved by varying the repetition rate of the time base via the diode circuit *V11* which limits the positive excursion of the anode voltage of *V8* and determines the end of the flyback. The variation of repetition rate, rather than the change of actual spot velocity, is essential in this time base since the time calibration holds good only with constant spot velocity.

Trigger

In the triggered condition of the time base, *V8* is connected as a transitron oscillator biased off on its suppressor and triggered from its control grid. The circuit comprises a phase-inverting tube *V12* and a buffer *V10*. The triggering pulse is fed to the TRIGGER OR SYNC terminal and to the grid of *V12* via *S16*, *C49*, and potentiometer *R120*. The polarity of the pulse fed to the grid of *V10* is selected by *S16*, which connects either the anode or cathode of *V12* to the buffer stage. The trigger signal is then injected into the control-grid circuit of *V8* via the anode of *V10*. By the control-grid modulation of *V8*, maximum sensitivity to the trigger pulse is achieved with complete isolation from the external pulse circuits by the interposition of the buffer stage.

PARTS LIST FOR COSSOR MODEL 1035

R1	8·2K	R26	82K	R48	3·9 MΩ	R74	3K	R96	¼ MΩ	R118	4·7K	C13	1 μF	C35	·015 μF
R2	2·2K	R27	82K	R49	22K	R75	47 Ω	R97	4·7K	R119	2·2K	C14	50 μF	C36	·05 μF
R3	82K	R28	27K	R50	2·2 MΩ	R76	120 Ω	R98	470K	R120	½ MΩ	C15	·25 μF	C37	·15 μF
R4	82K	R29	330 Ω	R51	82K	R77	200 Ω	R99	56K	R121	2·2K Ω	C16	8 μF	C38	·1 μF
R5	82K	R30	15K	R52	120 Ω	R78	15K	R100	27K	R122	5K Ω	C17	18 pF	C39	·005 μF
R6	82K	R31	82K	R53	100K	R79	15K	R101	120 Ω	R123	5K Ω	C18	5·6 pF	C40	1625 pF
R7	1·5K	R32	50K	R54	330K	R80	3·3 MΩ	R102	270 Ω	R124	330 Ω	C19	·1 μF	C41	450 pF
R8	33K	R33	1 MΩ	R55	1 MΩ	R81	47K	R103	10K	R125	470 Ω	C20	·25 μF	C42	115 pF
R9	68K	R34	1 MΩ	R58	1 MΩ	R82	146K	R104	10K	R126	15K Ω	C21	8 μF	C43	·5 μF
R10	680K	R35	100K	R59	2·2 MΩ	R83	120 Ω	R105	500 Ω	R127	15K Ω	C22	·25 μF	C44	·25 μF
R14	22K	R36	1 MΩ	R60	2·2 MΩ	R84	51K	R106	270K	C1	·25 μF	C23	·01 μF	C45	·005 μF
R15	50K	R37	1 MΩ	R63	13·2K	R85	14·5K	R107	47K	C2	16 μF	C24	1000 μF	C46	8 μF
R16	1 MΩ	R38	1 MΩ	R64	6·3K	R86	5·1K	R108	1000 Ω	C3	16 μF	C25	8 μF	C47	·1 μF
R17	2·2 MΩ	R39	1 MΩ	R65	1·5K	R87	2·2K	R109	180K	C4	32 μF	C26	·25 μF	C48	·25 μF
R18	2·2 MΩ	R40	2·2 MΩ	R66	600 Ω	R88	27K	R110	22K	C5	32 μF	C27	·1 μF	C49	·1 μF
R19	82K	R41	120 Ω	R67	150 Ω	R89	2·7K	R111	50K	C6	8 μF	C28	39 pF	C50	15 pF
R20	120K	R42	1 MΩ	R68	47 Ω	R90	1·5K	R112	22K	C7	·01 μF	C29	47 pF	C51	8 μF
R21	470K	R43	2·2K	R69	18 Ω	R91	1 MΩ	R113	56K	C8	·05 μF	C30	180 pF	C52	·25 μF
R22	100K	R44	82K	R70	2 Ω	R92	½ MΩ	R114	¼ MΩ	C9	·01 μF	C31	820 pF	C54	820 pF
R23	10K	R45	47 Ω	R71	470K	R93	120 Ω	R115	150K	C10	8 μF	C32	2200 pF	C55	6·8 pF
R24	50K	R46	120 Ω	R72	47K	R94	1 MΩ	R116	68K	C11	8 μF	C33	100 pF	C56	3·3 pF
R25	50K	R47	3·9 MΩ	R73	3K	R95	330 Ω	R117	4·7K	C12	1 μF	C34	·005 μF	C57	·1 μF

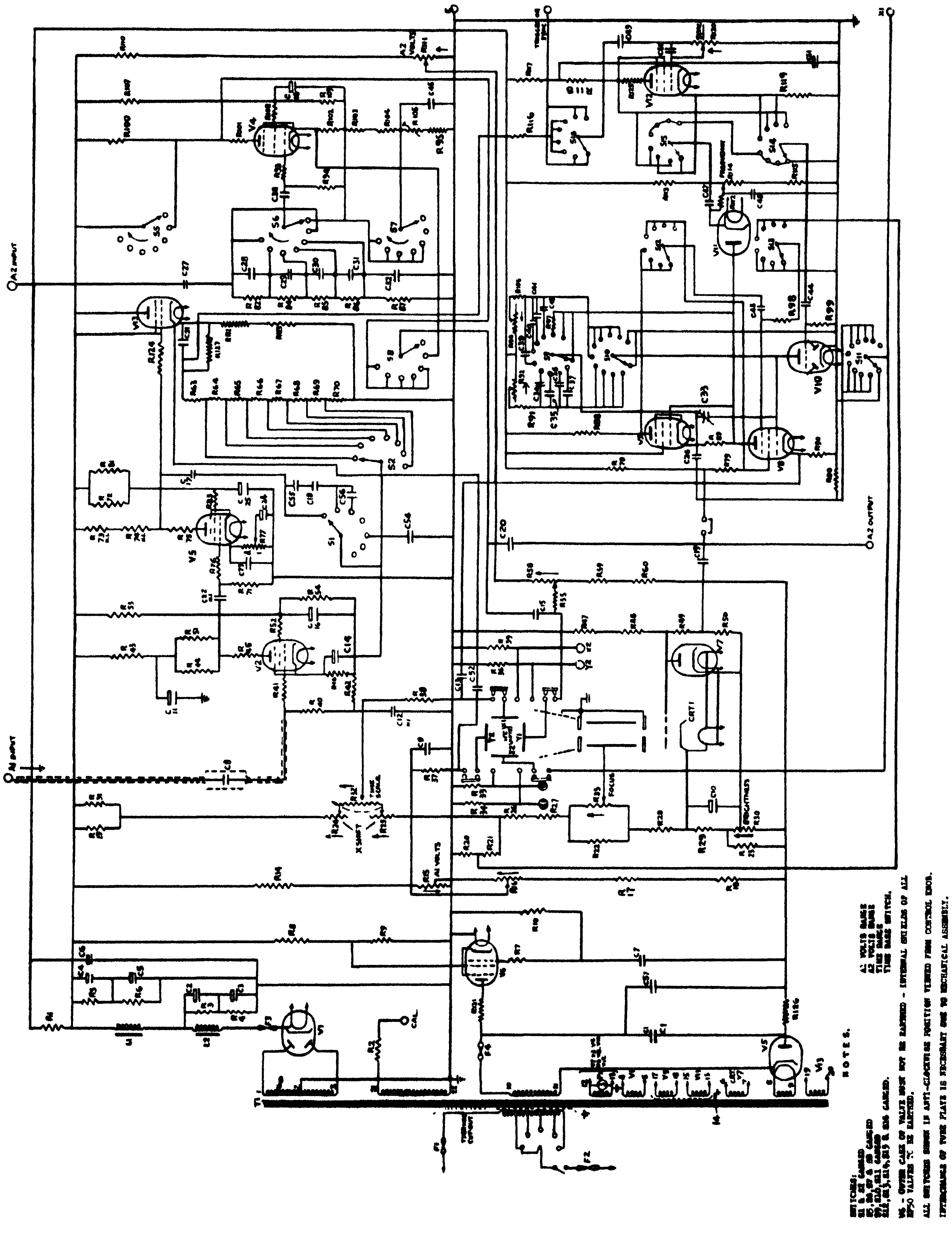

A2 INPUT
A1 INPUT
A2 OUTPUT
TIME BASE
FOCUS
BRILLIANCE
X SHIFT
A1 VOLTS
A2 VOLTS
CRT1
CAL.
NOTES.
SWITCHES:
S1 & S2 GANGED
S5, S6, S7 & S8 GANGED
S9, S10, S11 GANGED
S12, S13, S14, S15 & S16 GANGED.
A1 VOLTS RANGE
A2 VOLTS RANGE
TIME RANGE
TIME BASE SWITCH.
V6. - OUTER CASE OF VALVE MUST NOT BE EARTHED - INTERNAL SHIELDS OF ALL EF50 VALVES TO BE EARTHED.
ALL SWITCHES SHOWN IN ANTI-CLOCKWISE POSITION VIEWED FROM CONTROL KNOB.
INTERCHANGE OF TUBE PLATE IS NECESSARY DUE TO MECHANICAL ASSEMBLY.

Fig. 22-5.—Schematic of Cossor Model 1035.

Courtesy Cossor Ltd.

DU MONT MODEL 164-E

FREQUENCY RESPONSE
Vertical Amplifier 5 cps to 100,000 cps
Horizontal Amplifier 5 cps to 100,000 cps
Sweep Circuit 15 cps to 30,000 cps (sawtooth)

DEFLECTION FACTORS
Vertical Amplifier 0.80 rms volts/inch
Vertical-Deflection Plates 30 rms volts/inch
Horizontal Amplifier 0.65 rms volts/inch
Horizontal-Deflection Plates 30 rms volts/inch

LINE RATING 115/230 volts, 40-60 cps

TUBE COMPLEMENT

Type	*Function*
3AP (V1)	Cathode-Ray Tube
6C6 (V2)	Vertical Amplifier
6C6 (V3)	Horizontal Amplifier
Du Mont 2B4 (V4)	Sawtooth Sweep Oscillator
80 (V5)	Full-Wave Rectifier (Low Voltage)
80 (V6)	Half-Wave Rectifier (High Voltage)

The schematic circuit diagram for Model 164-E is shown in Fig. 22-6.

DU MONT MODEL 168

FREQUENCY RESPONSE
Vertical Amplifier 15 cps to 30,000 cps
Horizontal Amplifier 15 cps to 30,000 cps
Sweep Circuit 15 cps to 30,000 cps

DEFLECTION FACTORS
Vertical Amplifier 0.046 rms volts/inch
Vertical-Deflection Plates 20 rms volts/inch

LINE RATING 115/230 volts, 40-60 cps

TUBE COMPLEMENT

Type	*Function*
6C6 (V1)	Vertical Amplifier
76 (V2)	Vertical Amplifier
5MP (V3)	Cathode-Ray Tube
Du Mont 2B4 (V4)	Sawtooth Sweep Oscillator
6C6 (V5)	Horizontal Amplifier
80 (V6)	Full-Wave Rectifier (Low Voltage)
80 (V7)	Half-Wave Rectifier (High Voltage)

The schematic circuit diagram of Model 168 is shown in Fig. 22-7. A special 3-position switch enables the input signal to be applied directly to the plates of the cathode-ray tube, through the first amplifier alone, or through both vertical amplifiers.

PARTS LIST FOR DU MONT MODEL 164-E

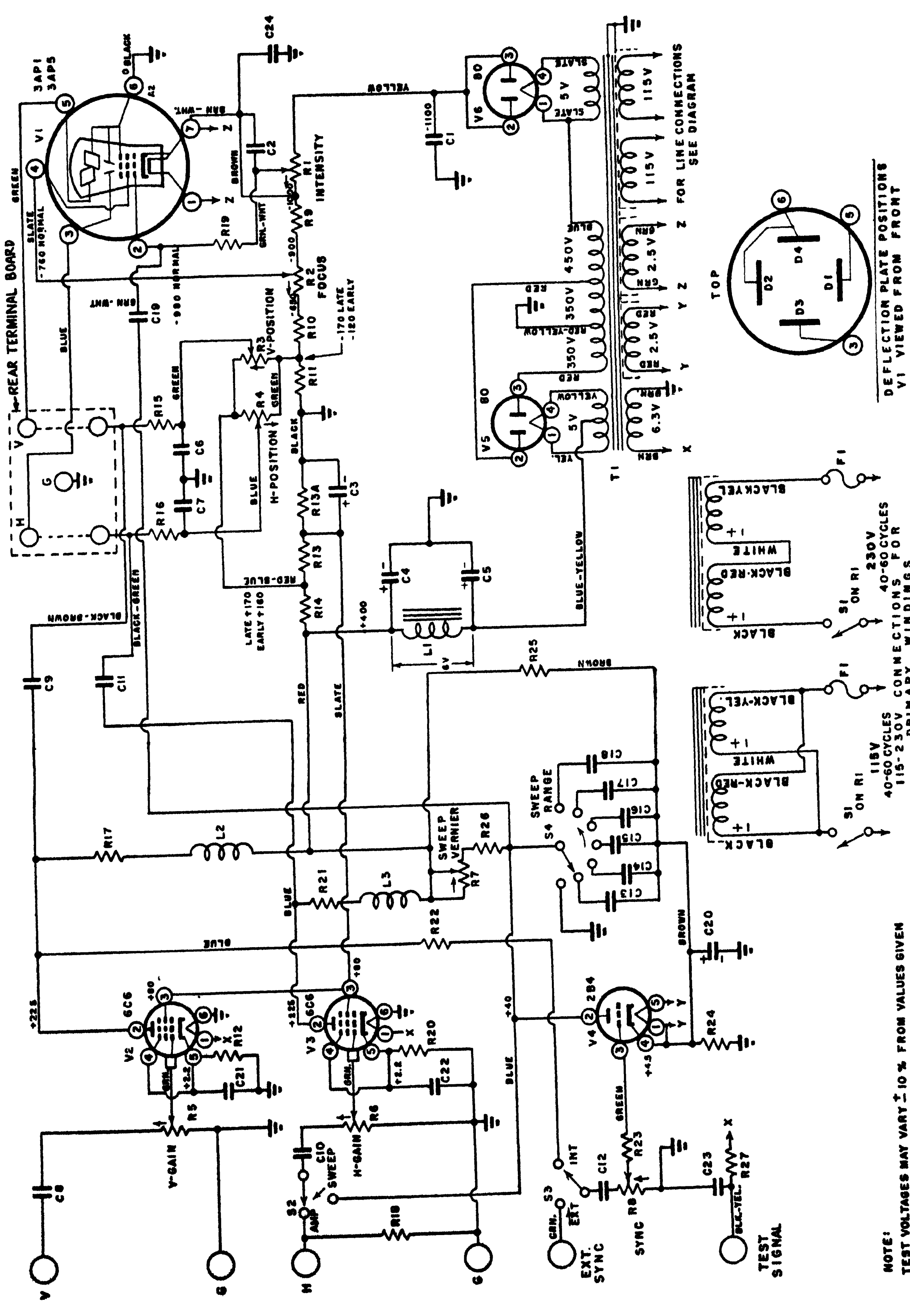
C1	0.5 µf 1500V
C2	0.5 µf 600V
C3	8 µf 150V elec.
C4 C5	4+4 µf 475V elec.
C6	0.05 µf 400V
C7	" "
C8	0.25 µf 400V
C9	" "
C10	" "
C11	" "
C12	0.05 µf 400V
C13	0.2 µf "
C14	0.04 µf "
C15	0.01 µf "
C16	2400 µµf 500V
C17	620 µµf "
C18	120 µµf "
C19	50 µµf 1200V
C20	25 µf 50V elec.
C21	3900 µµf 500V
C22	3900 µµf "
C23	0.1 µf 1000V
C24	0.1 µf 1600V
L1	8 h 0.035 amp
L2	60 mh Peaking Coil
L3	60 mh " "

R1	200 K ½W
R2	500 K ½W
R3	4 meg. 2 W
R4	" "
R5	1 meg. ½W
R6	4 meg. 2 W
R7	" "
R8	15 K ½W
R9	100 K ½W ±10%
R10	560 K 1W ±10%
R11	390 K 1W ±10%
R12	820 ohm ½W "
R13	10 K 3W "
R13A	" " "
R14	25 K 10W ±5%
R15	4.7 meg. ½W ±10%
R16	4.7 meg. ½W ±10%
R17	91 K 1W ±5%
R18	1 meg. ½W ±10%
R19	470 K ½W "
R20	820 ohm ½W "
R21	91 K 1W ±5%
R22	1 meg. ½W ±10%
R23	100 K ½W "
R24	1 K ½W "
R25	100 K 3W "
R26	470 K 1W "
R27	10 K ½W "

PARTS LIST FOR DU MONT MODEL 168

C1	1µf. 400v.
C2	0.1µf. 1000v.
C3	0.1µf. 1000v.
C4	0.05µf. 400v.
C5	0.1µf. 1000v.
C6	200µµf. 500v.
C7	600µµf. 500v.
C8	2500µµf. 500v.
C9	0.01µf. 400v.
C10	0.04µf. 400v.
C11	0.2µf. 400v.
C12	1µf. 400v.
C13	0.05µf. 400v.
C14	0.05µf. 400v.
C15	0.05µf. 400v.
C16	8µf. 150v.
C17	4µf. 450v.
C18	4µf. 450v.
C19	0.5µf. 600v.
C20	0.5µf. 1500v.
C21	50µµf. 1200v.
C22	4µf. 450v.
C23	4µf. 450v.
C24	0.1µf. 1600v.
C25	25µf. 50v.
F1	1 amp. 250v.
L1	8.5H. 35ma. 325 ohms d.c
L2	8.5H. 35ma. 325 ohms d.c
R1	1meg. pot.

R2	4meg. pot.
R3	15K pot.
R4	4meg. pot.
R5	4meg. pot.
R6	4meg. pot.
R7	500k pot.
R8	200K pot.
R9	100K 1 W.
R10	500K ½ W.
R11	6K ½ W.
R12	100K 1 W.
R13	1K ½ W.
R14	100K 1 W.
R15	1meg. ½ W.
R16	100K ½ W.
R17	750 ohms ½ W.
R18	750K. 1 W.
R19	100K. 3 W.
R20	10K. ½ W.
R21	5meg. ½ W.
R22	1meg. ½ W.
R23	25K. 10 W.
R24	10K. 3 W.
R25	10K. 3 W.
R26	220 ohms ½ W.
R27	250K. ½ W.
R28	750K. 1 W.
R29	100K. ½ W.
R30	1meg. ½ W.
R31	3K. ½ W.
R32	3K ½ W.



Courtesy Du Mont Labs.

NOTE: TEST VOLTAGES MAY VARY ± 10% FROM VALUES GIVEN

Fig. 22-6.—Schematic of Du Mont Model 164-E.

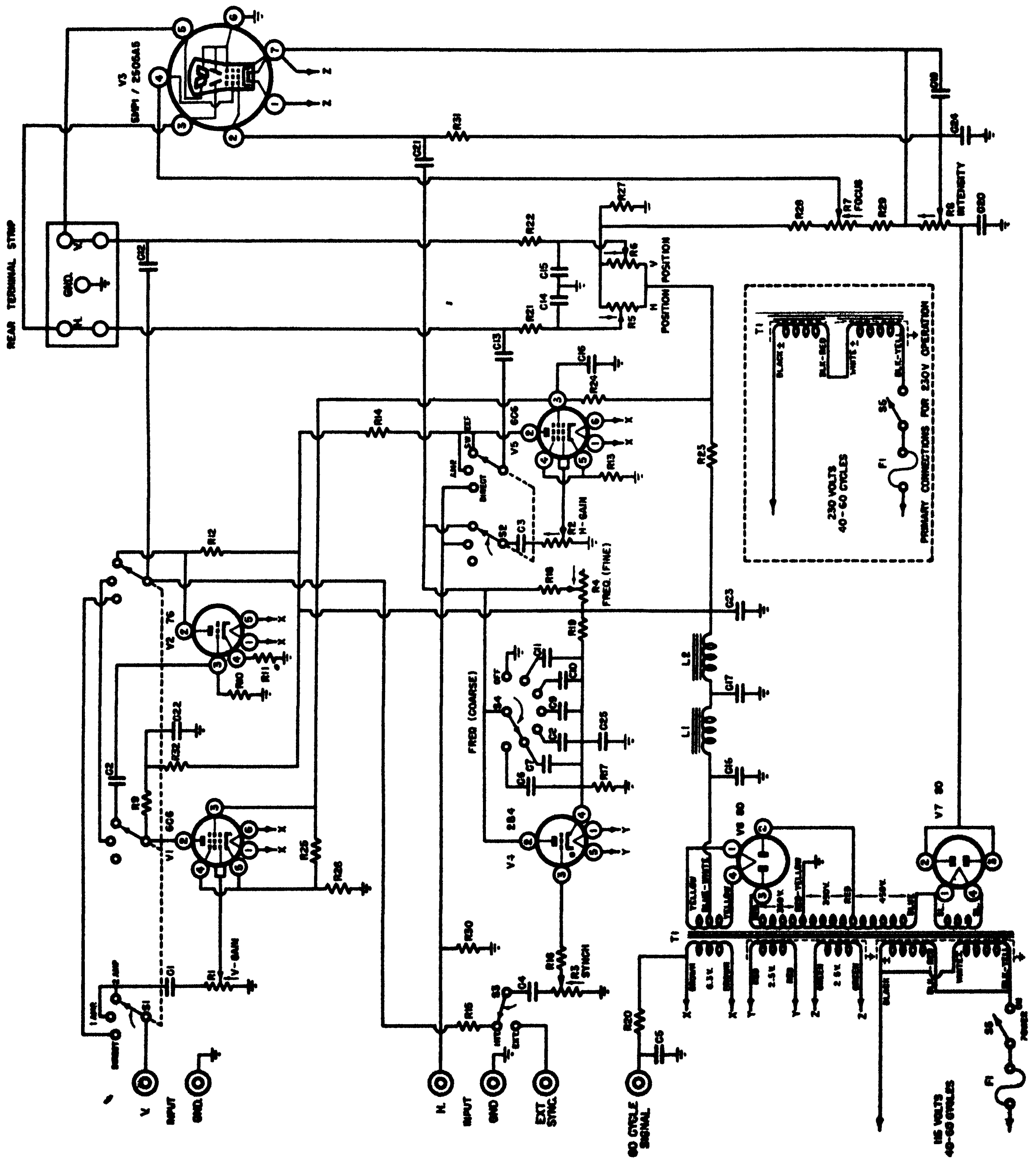


Fig. 22-7.—Schematic of Du Mont Model 168.

Courtesy Du Mont Labs.

DU MONT MODEL 175-A

FREQUENCY RESPONSE
Vertical Amplifier 5 cps to 100,000 cps, ±10%
Horizontal Amplifier 5 cps to 100,000 cps, ±10%
Sweep Circuit 15 cps to 30,000 cps

DEFLECTION FACTORS (Maximum)
Vertical Amplifier 0.030 rms volts/inch
Vertical-Deflection Plates 24.5 rms volts/inch
Horizontal Amplifier 0.033 rms volts/inch
Horizontal-Deflection Plates 27.0 rms volts/inch

LINE RATING 115/230 volts, 40-60 cps

TUBE COMPLEMENT

Type	*Function*
6SJ7 (V1)	Vertical Amplifier
6K6G (V2, V3)	Vertical Output Amplifiers
6SJ7 (V4)	Horizontal Amplifier
6K6G (V5, V6)	Horizontal Output Amplifiers
Du Mont 2B4 (V7)	Sawtooth Sweep Oscillator
6J5G (V8)	Sweep Amplifier
6J7G (V9)	Z-Axis Amplifier
80 (V10)	Rectifier, Negative Supply
2X2 (V11)	Rectifier, High-Voltage
5Z3 (V12)	Rectifier, Low-Voltage
5LP (V13)	Cathode-Ray Tube
6H6 (V14)	Single Sweep Control
6F8G (V15)	Voltage Regulator

Vertical Input

The schematic circuit diagram of Model 175-A is shown in Fig. 22-8. There is a 5-pole 4-position switch in the vertical section of the oscilloscope. Two of the positions apply the input signal dlirectly to the plates of the cathode-ray tube, and in the other two positions, the input signal passes through the vertical amplifiers. In both the direct or amplifier sections of this switch there is an *L* or *H* position. The *L* positions provide a high input impedance of one megohm and are primarily for low-frequency signal inputs. The *H* positions provide a low input impedance of 100,000 ohms and are primarily used for high-frequency inputs. In either case the signal input is capacitance coupled to the amplifiers or directly to the cathode-ray-tube plates through a 0.5-µf capacitor.

Sweep Circuit

A special diode circuit is incorporated in the sweep network to provide single sweep control of the sawtooth oscillations. One-half of a 6H6 duo-diode tube is used for this purpose. The plate of the diode is connected to the grid circuit of the 2B4 tube and the cathode of the diode grounded through a 50,000-ohm potentiometer and a 20,000-ohm resistor. The cathode of the diode is connected directly to the variable arm of the potentiometer. In the normal operation of the sweep circuit, the cathode of the diode is positive with respect to its plate so that the diode will be nonconducting and thus will not have any effect on the circuit. The normal plate voltage on the 2B4 tube is 32 volts, which means that this same voltage appears on the plate of the diode. The cathode voltage on the diode varies from a low of +18 volts to a high of +60 volts as determined by the setting of the potentiometer. This means that the diode will conduct only when the cathode is +32 volts or less. The diode will, therefore, start conducting just below the breakdown potential of the discharge tube. This action prevents the charging capacitor in the sweep circuit from ever attaining a potential high enough for sawtooth oscillations to take place.

When a positive synchronizing signal is applied to the grid of the discharge tube, the breakdown potential is reduced, the diode does not conduct, the capacitor discharges through the discharge tube and then recharges again to the value limited by the voltage on the plate of the diode. This produces a single linear sweep of the beam across the cathode-ray-tube screen because the diode starts conducting again when the capacitor recharges to its full value. Continued oscillations of the sweep are prevented because the diode is always conducting. When another synchronizing signal is applied to the grid of the discharge tube, the breakdown potential of the tube will again be lowered so that another single linear sweep occurs.

Grid Modulation of Intensity

This oscilloscope has special provision for grid modulation of intensity. In order to accomplish this, a 6J7G amplifier, called the Z-axis amplifier, is employed to amplify any input signal applied to the Z-axis terminals of the unit.

Blanking

The Z-axis amplifier is also used for blanking the beam of the cathode-ray tube during the retrace time of the sweep signal. This is accomplished by connecting the plate circuit of the sweep amplifier to the grid circuit of the Z-axis amplifier through a high-pass filter network differentiating circuit, which basically consists of a 50-µµf capacitor and a 50,000-ohm resistor. When the sawtooth wave passes through this filter network, the signal applied to the grid of the Z-axis amplifier is in the form of a pulse. The width of the pulse is approximately the same as the time duration of the sawtooth retrace and is considered as a direct function of the retrace time. Since differentiation always intensifies rapid amplitude changes, it is apparent that the pulse input to the Z-axis amplifier is primarily due to the retrace of the sawtooth wave. This pulse is then amplified and applied to the grid circuit in such a manner that it will bias the grid to the point of cutoff and hence blank out the return trace of the spot on the screen.

PARTS LIST FOR DU MONT MODEL 175

R1	1 meg. section
R2	100 K section
R3	500 ohms ½ watt
R4	50 K 1 watt
R5	50 K 1 watt
R6	25 K ½ watt
R7	2 meg. ½ watt
R8	1250 ohms ½ watt
R9	12 K 10 watt
R10	850 ohms ½ watt
R11	2 meg. ½ watt
R12	13.5 K 10 watt
R13	2.5 K 10 watt
R14	200 K ½ watt
R15	1 meg. ½ watt
R16, R16A	1.5 meg.
R17	1 meg. ½ watt
R18	5 meg. ½ watt
R19	1 meg. pot.
R20	500 ohms ½ watt
R21	50 K 1 watt
R22	50 K 1 watt
R23	25 K ½ watt
R24	2 meg. ½ watt
R25	1250 ohms ½ watt
R26	12 K 10 watt
R27	850 ohms ½ watt
R28	2 meg. ½ watt
R29	13.5 K 10 watt
R30	2.5 K 10 watt
R31	1 meg. ½ watt
R32, R32A	1.5 meg.
R33	1 meg. ½ watt
R34	25 K pot.
R35	100 K ½ watt
R36	1250 ohms ½ watt
R37	100 K 3 watt
R38	750 K 1 watt
R39	4 meg. pot. ±10%
R40	25 K ½ watt
R41	50 K ½ watt
R42	20 K ½ watt
R43	1 meg. pot.
R44	300 ohms ½ watt
R45	15 K 3 watt
R46	30 K 10 watt
R47	25 K ½ watt
R48	10 K ½ watt
R49	10 K ½ watt
R51	50 K ½ watt
R52	50 K ½ watt
R53	10 meg. ½ watt
R54	10 meg. ½ watt
R55	100 K pot.
R56	100 K ½ watt
R57	100 K ½ watt
R58	500 K pot.
R59	750 K 1 watt
R60	10 K ½ watt
R61	50 K ½ watt
R62	150 ohms ½ watt
R63	75 K ½ watt
R64	50 K ½ watt
R65	20 K ½ watt
R66	50 K pot.
R67	250 K ½ watt
R68	1 meg. ½ watt
R69	150 K ½ watt
R70	35 K ½ watt
R71	1 meg. ½ watt
R72	10 K ½ watt
R73	25 K ½ watt
R74	5 meg. ½ watt
R75	2 K ½ watt
R76	2 K ½ watt
R77	50 K ½ watt
R78	50 K ½ watt
C1	0.5 µf. 600 W.V.
C2	4 µf. 475 W.V.
C3	0.5 µf. 600 W.V.
C4	0.5 µf. 600 W.V.
C5	4 µf. 475 W.V.
C6	0.25 µf. 400 W.V.
C7	0.25 µf. 400 W.V.
C8	0.05 µf. 400 W.V.
C9	0.05 µf. 400 W.V.
C10	0.05 µf. 400 W.V.
C11	0.5 µf. 600 W.V.
C12	4 µf. 475 W.V.
C13	0.5 µf. 600 W.V.
C14	0.5 µf. 600 W.V.
C15	4 µf. 475 W.V.
C16	0.25 µf. 400 W.V.
C17	0.25 µf. 400 W.V.
C18	0.05 µf. 400 W.V.
C19	0.05 µf. 400 W.V.
C20	0.25 µf. 400 W.V.
C21	0.25 µf. 400 W.V.
C22	0.05 µf. 400 W.V.
C23	0.01 µf. 400 W.V.
C24	0.0025 µf. 500 W.V.
C25	500 µµf. 500 W.V.
C26	100 µµf. 500 W.V.
C28	50 µµf. 500 W.V.
C29	8 µf. 450 W.V.
C30	0.05 µf. 400 W.V.
C31	4 µf. 475 W.V.
C32	4 µf. 475 W.V.
C33	0.5 µf. 1600 W.V.
C35	0.5 µf. 600 W.V.
C36	0.05 µf. 1600 W.V.
C37	0.1 µf. 1600 W.V.
C38	0.5 µf. 1500 W.V.
C39	8 µf. 450 W.V.
C40	8 µf. 450 W.V.
C41	8 µf. 450 W.V.
C42	0.05 µf. 400 W.V.
C43	16 µf. 450 W.V.
C44	0.5 µf. 1500 W.V.
C45	0.05 µf. 400 W.V.
C46	0.003 µf. 500 W.V.
C47	0.003 µf. 500 W.V.
C48	2 or 3 or 4 µf. 450 W.V.
C49	0.25 µf. 400 W.V.
C50	4 µf. 475 W.V.
C51	4 µf. 475 W.V.

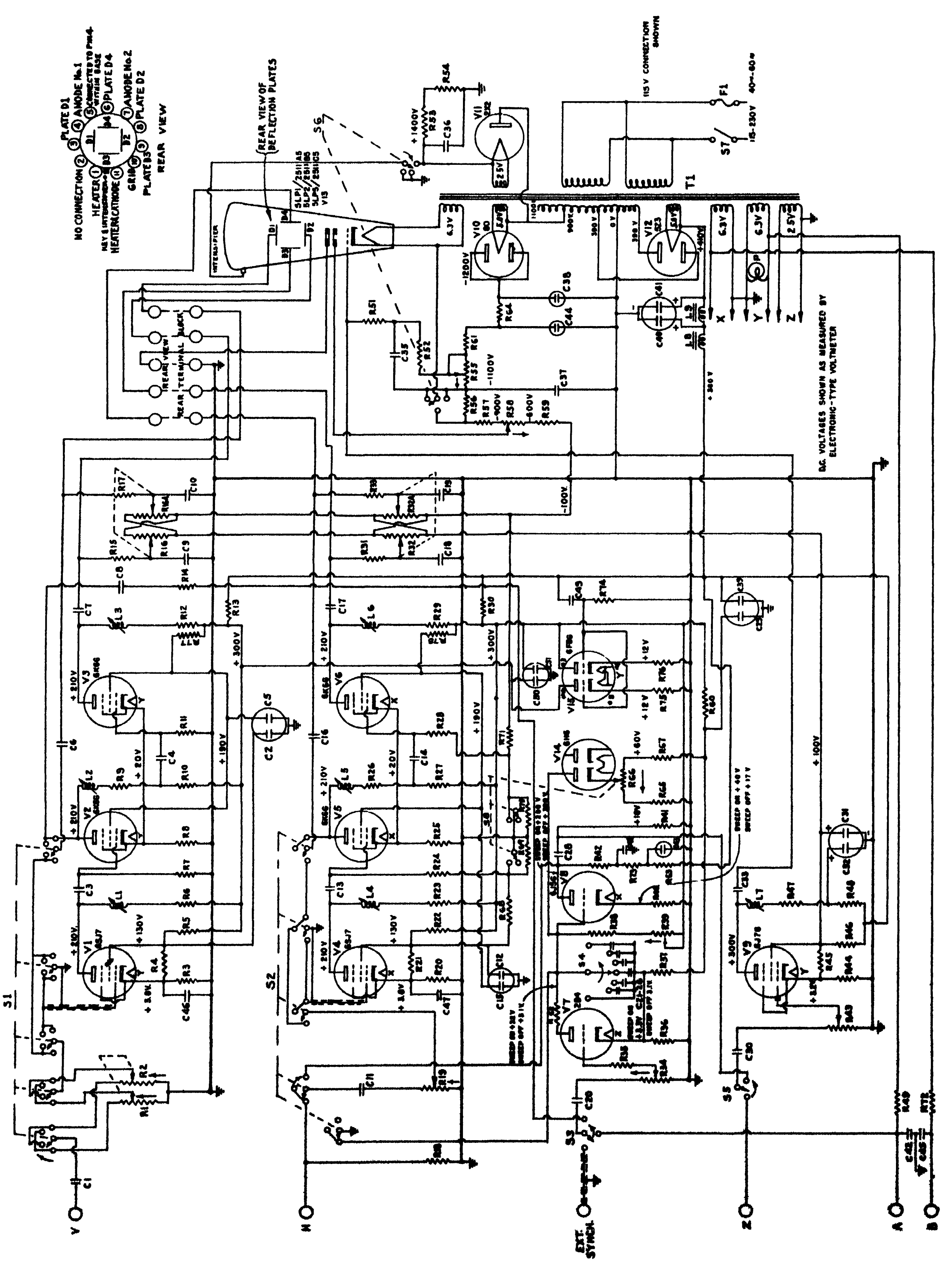


Fig. 22-8.—Schematic of Du Mont Model 175-A.

Courtesy Du Mont Labs.

DU MONT MODELS 208 AND 208-B

FREQUENCY RESPONSE
Vertical Amplifier 2 cps to 100,000 cps, ±10%
Horizontal Amplifier 2 cps to 100,000 cps, ±10%
Sweep Circuit 2 cps to 50,000 cps

DEFLECTION FACTORS
Vertical Amplifier 0.010 rms volts/inch
Vertical-Deflection Plates 21 rms volts/inch
Horizontal Amplifier 0.5 rms volts/inch
Horizontal-Deflection Plates 22 rms volts/inch

LINE RATING 115-230 volts, 40-60 cps

TUBE COMPLEMENT

Type	*Function*
6SN7GT (V1, V4)	Impedance Transformation Y-Axis Input Stage, Y-Axis Positioning Amplifier
6SN7GT (V2, V3)	Y-Axis Amplifier
6V6 (V5, V6)	Y-Axis D-C Coupled Phase-Inversion Amplifiers
6Q5G (V7)	Sawtooth Sweep Oscillator
6SN7GT (V8, V9)	Impedance Transformation X-Axis Input Stage, X-Axis Positioning Amplifier
6V6 (V10, V11)	X-Axis D-C Coupled Phase-Inversion Amplifiers
5LP-A (V12)	Cathode-Ray Tube
80 (V13)	High-Voltage Half-Wave Rectifier
80 (V14)	Low-Voltage Full-Wave Rectifier
6X5GT/G (V15)	Half-Wave Negative Supply Rectifier
6V6 (V16)	Series-Voltage Regulator
6SJ7 (V17)	Voltage-Regulator Control Tube
991 (V18)	Neon Tube Voltage Regulator

The schematic circuit diagram of Model 208-B is shown in Fig. 22-9. The only differences between Models 208 and 208-B are that 6SN7GT tubes are used in Model 208-B, where Model 208 uses type 6F8G tubes instead; also the 5LP- series cathode-ray tubes in Model 208 have been replaced with 5LP-A series in Model 208-B which are tubes of more recent design. The analysis to follow will be with respect to Model 208-B. However, whatever is said regarding the circuit analysis of Model 208-B will also hold true for Model 208.

Input Circuits

A two-step attenuator is used in the vertical input section. This attenuator, whether in the "Under 25 volt rms" input position or "Under 250 volt rms" input position, controlled by switch *S1* as shown on the schematic, limits the signal voltage input to the first stage to 25 volts rms or less. When switch *S1* is in the "Under 250 volt rms" position, a voltage-divider network is inserted in the input section and the signal voltage splits up in such a manner that the voltage drop across the input grid circuit, consisting of *R2*, *R3*, and *C3* all in parallel, is about 10 per cent of the input signal. The remaining 90 per cent of the signal voltage appears across the parallel circuit of *R1* and *C2*. This input switch attenuator is employed to prevent overloading of the input stage. The input impedance is essentially the same for each position of the switch.

The input stage in the vertical section is in the form of an impedance transforming circuit. This arrangement permits the use of a low-impedance vertical gain control with a minimum amount of frequency discrimination and at the same time provides for a high-impedance input. This input stage is a cathode-follower circuit which utilizes one triode section *V1* of a 6SN7GT tube.

Positioning Control

The output from *V3* is fed to the grid circuit of *V4* which is a "positioning amplifier." This amplifier is also arranged in the form of a cathode-follower circuit, with the signal appearing across the cathode load. In this cathode circuit, the 15,000-ohm potentiometer *R16* which functions as the positioning control is in series with a 300,000-ohm resistor; the total signal voltage appearing across these two resistors.

With a d-c potential of —280 volts at the bottom of *R17* and the direct current of *V4* flowing through the cathode circuit, the potential across *R16* is such that the cathode side of this resistor is 7.2 volts positive as compared to a negative voltage of —7.1 volts at its other end. This means that, at approximately the center of *R16*, the d-c voltage is zero. Thus we can understand how this potentiometer can function as a positioning control. The variable arm of *R16* is coupled directly to the grid of *V5*, a 6V6 tube acting as part of a direct-coupled phase-inverter circuit. Since the coupling is direct, both the signal and d-c voltages in the cathode load circuit of *V4* are fed to the grid of *V5*. The 15,000-ohm value of *R16* is small compared to that of *R17*, hence the change in signal voltage fed to *V5*, as *R16* is varied, is considered to be negligible. This variation of *R16* effectively controls only the d-c level of voltage input to the grid of *V5*. Since the signal voltage as well as the positioning voltage are applied together to the grid of *V5*, the signal voltage will move the screen spot about whatever positive value it assumes. The X-axis amplifier circuit is essentially the same as the Y-axis except that a two-stage amplifier is not employed after the impedance transformation stage.

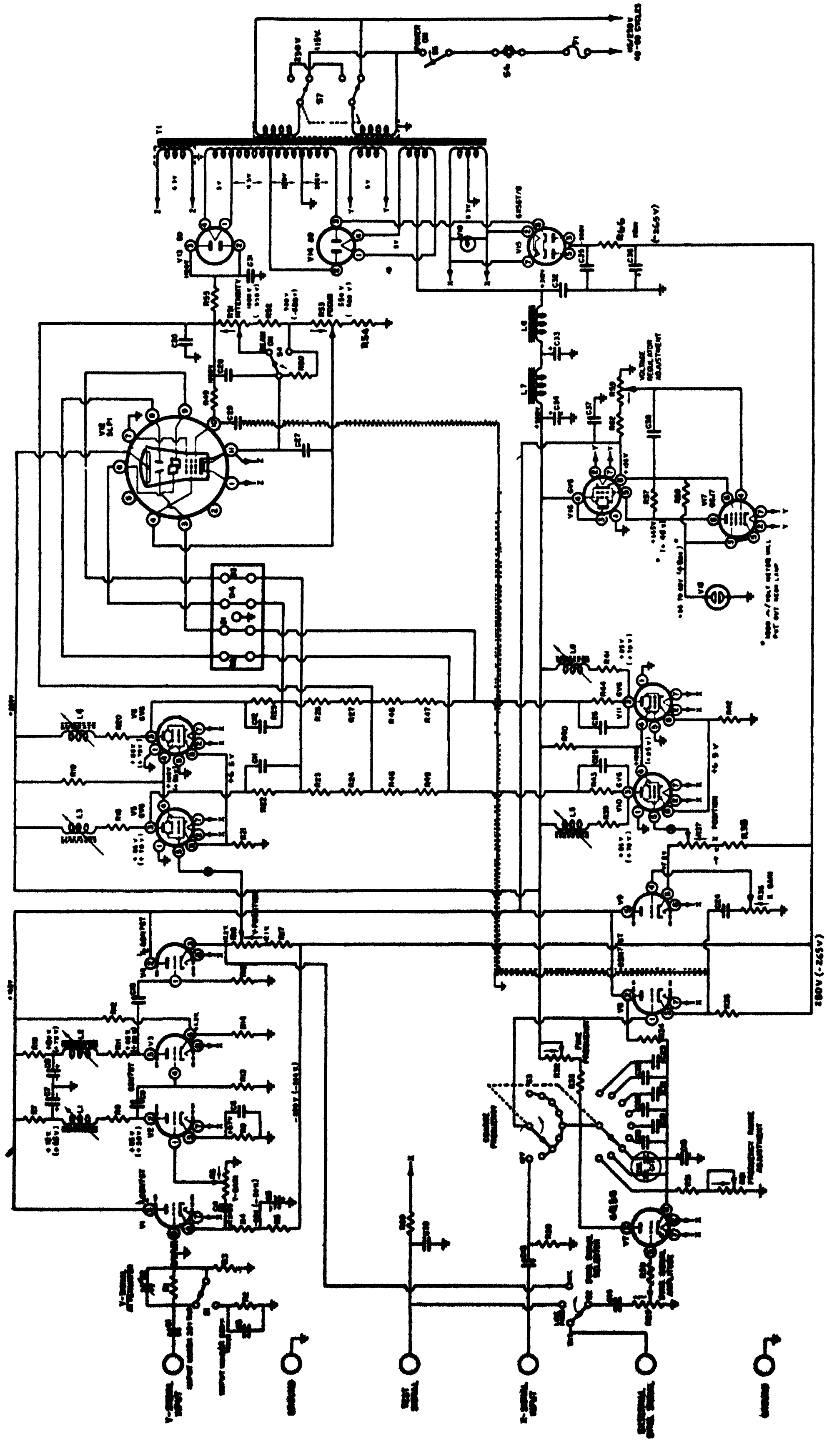
PARTS LIST FOR DU MONT MODELS 208 AND 208-B

C1—0.5μf. 600V.
C2—4-30μμf. trimmer
C3—200μμf. 500V.
C4—8μf. 200V.
C5—24μf. 350V. elec.
C6—0.005μf. 500V.
C7—30μf. 150V. elec.
C8—30μf. 150V. elec.
C7 and C8 common neg.
C9—1μf. 200V.
C10—1μf. 200V.
C11—0.25μf. 400V.
C12—0.25μf. 400V.
C13—0.25μf. 400V.
C14—0.1μf. 1000V.
C15—25μf. 50V. elec.
C16—1μf. 400V.
C17—0.2μf. 400V.
C16-C17 common can
C18—0.04μf. 400V.
C19—0.01μf. 400V.
C20—2500μμf. 500V.
C21—600μμf. 500V.
C22—125μμf. 500V.
C23—70μμf. 500V.
C24—8μf. 200V.
C25—0.25μf. 400V.
C26—0.25μf. 400V.
C27—0.1μf. 1000V.
C28—0.1μf. 1000V.
C29—50μμf. 1200V.
C30—0.5μf. 1500V.
C31—0.5μf. 1500V.
C32—1μf. 1000V.
C33—16μf. 450V. elec.
C34—40μf. 450V. elec.
C35—1μf. 1000V.
C36—16μf. 450V.
C37—1μf. 200V.
C38—1μf. 200V.
C39—0.1μf. 1000V.

F1—1.5 amps. fuse

L1—1-3.1mh.
L2—1-3.1mh.
L3—7-19mh.
L4—7-19mh.
L5—7-19mh.
L6—7-19mh.
L7—8h 325 ohms d.c.
L8—8h 325 ohms d.c.

R1—2meg. ½W.±5%
R2—250K ½W.±5%
R3—2meg. ½W.
R4—100K 1W.
R5—250K 1W.
R6—100K pot.
R7—25K 1W.
R8—8K 1W.
R9—250 ohm ½W.
R10—25K 1W.
R11—8K 1W.
R12—50K 1W.
R13—1meg. ½W.
R14—250 ohm ½W.
R15—1meg. ½W.
R16—15K pot.
R17—300K 1W.
R18—25K 10W.
R19—150K 1W.
R20—25K 10W.
R21—400 ohm 1W.
R22—750K 1W.
R23—5meg. 1W.
R24—5 meg. 1W.
R25—750K 1W.
R26—5meg. 1W.
R27—5meg. 1W.
R28—5meg. ½W.
R29—100K pot.
R30—10K ½W.
R31—500 ohm ½W.
R32—5meg. pot.
R33—500K 1W.
R34—39K 3W.
R35—250K 1W.
R36—500K pot.
R37—15K pot.
R38—300K 1W.
R39—25K 10W.
R40—150K 1W.
R41—25K 10W.
R42—400 ohm 1W.
R43—750K 1W.
R44—750K 1W.
R45—5meg. 1W.
R46—5meg. 1W
R47—5meg. 1W.
R48—5meg. 1W.
R49—100K ½W.
R50—1meg. ½W.
R51—100K pot.
R52—200K 1W.
R53—500K pot.
R54—1meg. 1W.
R55—50K 1W.
R56—47K 3W.
R57—500K 1W.
R58—100K 1W.
R59—500K pot.
R60—10K ½W.
R61—1K pot.
R62—500K 1W.



Courtesy Du Mont Labs

Fig. 22-9.—Schematic of Du Mont Model 208-B.

• DU MONT MODEL 224-A

Frequency Response

Vertical Amplifier 20 cps to 2 Mc (unifrom within 3 db)
Horizontal Amplifier 10 cps to 100 kc (uniform within 3 db)
Sweep Circuit 15 cps to 30 kc

Deflection Factors

Vertical Amplifier 0.1 rms volt/inch
Vertical-Deflection Plates 25 rms volts/inch
Horizontal Amplifier 0.7 rms volt/inch
Horizontal-Deflection Plates 28 rms volts/inch

Line Rating 115 volts, 60 cps

Tube Complement

Type	*Function*
6J5 (V1)	Vertical Amplifier
6AC7 (V2)	Vertical Amplifier
6AG7 (V3, V4)	Dual Vertical Amplifiers
3GP1 (V5)	Cathode-Ray Tube
6SN7 (V6)	Synchronizing and Horizontal Amplifiers
6Q5G (V7)	Horizontal Amplifier
6SG7 (V8, V9)	Dual Horizontal Amplifiers
5Z3 (V10)	Full-Wave Rectifier
80 (V11)	Half-Wave Rectifier
6V6GT (V12)	Voltage Control
6SJ7 (V13)	Voltage Regulator
991 (V14)	Neon Tube

The schematic circuit diagram of Model 224-A is shown in Fig. 22-10. Switch *S1* controls the attenuation of the vertical input, using resistive-capacitive elements which vary the impedance from *C1* to ground and, therefore, the input impedance of *V1* when *S2* is set for vertical input *Y* on the diagram.

S2 switches *V1* from the Y-input to the probe connector, or vice versa, as desired.

The power supply uses a straightforward circuit with the possible exception of the voltage-regulator system which is somewhat different to merit some special comment. Essentially, the 6V6GT *V12* is a variable plate-cathode resistance in series with *R59* and *R56*. Therefore, the voltage drops of these resistors are a function of the 6V6GT plate-cathode resistance which, in turn, is controlled by the source voltage and the control tube, a 6SJ7, marked *V13* on the diagram.

The *V12* screen and *V13* plate are connected to the +190-volt point in the circuit. This connection is made through *R71*. If the voltage across *C43* rises for any reason, the voltages across *R70* and *R69* increase and the bias on *V13* rises. The reduced current in *R71* causes an increased screen potential for *V12* and effectively reduces the plate current, causing in effect an increased plate resistance and a constant voltage drop effect in *V12* is obtained. If the supply voltage across *C43* drops, the resistance of *V12* goes down and, in any case, the potential across *C49* is a steady 190 volts d.c. Neon tube *V14* stabilizes the potential drop across that portion of *R69* connected between *C50* and ground. The voltage across *C50* is the control voltage for *V13*.

PARTS LIST FOR DU MONT MODEL 224-A

C1	0.2 μf. 400V.
C2	3-12 μμf. 500V.
C3	3-12 μμf. 500V.
C4	.001 μf. 500V.
C5	70 μμf. 500V.
C6	0.2 μf. 400V.
C7	100 μf. 50V. elec.
C8	0.1 μf. 1000V.
C9	0.5 μf. 600V.
C10	0.5 μf. 600V.
C11	4 μf. 600V.
C12	0.5 μf. 200V.
C13	0.5 μf. 600V.
C14	25 μf. 50V. elec.
C15	0.1 μf. 1000V.
C16	0.1 μf. 1000V.
C17	0.25 μf. 400V.
C18	.05 μf. 400V.
C19	0.1 μf. 1000V.
C20	0.1 μf. 1000V.
C21	0.5 μf. 600V.
C22	0.25 μf. 400V.
C23	70 μμf. 500V.
C24	3-12 μμf. 500V.
C25	100 μf. 50V. elec.
C26	0.2 μf. 400V.
C27	0.5 μf. 600V.
C28	3-12 μμf. 500V.
C29	0.15 μf. 400V.
C30	.04 μf. 400V.
C31	.01 μf. 400V.
C32	2500 μμf. 500V.
C33	600 μμf. 500V.
C34	150 μμf. 500V.
C35	25 μf. 50V. elec.
C36	0.5 μf. 600V.
C37	0.1 μf. 1000V.
C38	0.1 μf. 1000V.
C39	Deleted
C40	Deleted
C41	Deleted
C42	Deleted
C43	4 μf. 600V.
C44	4 μf. 600V.
C45	4 μf. 600V.
C46	0.5 μf. 1500V.
C47	0.5 μf. 1500V.
C48	0.5 μf. 600V.
C49	0.5 μf. 600V.
C50	0.5 μf. 600V.
C51	.05 μf. 400V.
C52	4-30 μμf. 500V.
C53	.05 μf. 1600V.
F1	3 amp. fuse
L1	70-250 μh.
L2	170 μh.
L3	170 μh.
L4	8.5 mh.
L5	8.5 mh.
L6	19 h. 150 ma.
R1	2 meg. ½W.±5%
R2	2 meg. ½W.±5%
R3	20 K ½W.±5%
R4	250 K ½W.±5%
R5	2 meg. ½W.
R6	50 ohm ½W.
R7	1 K ½W.
R8	1 K pot.
R9	110 ohm ½W.
R10	25 meg. 1W.
R11	75 K 1W.
R12	100 K ½W.
R13	2.5 K 1W.
R14	10 K 3W.
R15	50 ohm ½W.
R16	75 K ½W.
R17	100 K 1W.
R18	15 K ½W.
R19	470 ohm 3W.
R20	50 ohm ½W.
R21	3.5 K 10W. non ind.
R22	15 K 10W.
R23	3.5 K 10W. non ind.
R24	35 K ½W.
R25	500 K ½W.
R26	1.5 K ½W.
R27	8 K ½W.
R28	200 K C.T. pot.
R29	20 K ½W.
R30	10 K 1W.
R31	10 K 1W.
R32	1 K ½W.
R33	500 K 1W.
R34	4 meg. pot.
R35	250 K ½W.±5%
R36	2 meg. ½W.±5%
R37	2 meg. ½W.
R38	10 K ½W.
R39	10 K pot.
R40	Deleted
R41	3 meg. ½W.
R42	82K 3W.
R43	100 K 1W.
R44	50 ohm ½W.
R45	15 K ½W.
R46	2 K 1W.
R47	2 meg. ½W.
R48	50 ohm. ½W.
R49	82K 3W.
*R50	37.5 K 2W.
R51	20 K ½W.
R52	20 K ½W.
*R53	37.5 K 2W.
R54	5 meg. ½W.
R55	4 meg. dual pot.
R56	5 meg. ½W.
R57	5 meg. ½W.
R58	5 meg. ½W.
R59	100 K 1W.
R60	100 K 1W.
R61	4 meg. dual pot.
R62	10 K 1W.
R63	100 K. 1W.
R64	100 K pot.
R65	150 K 1W.
R66	500 K pot.
R67	750 K 1W.
R68	200 K 1W.
R69	500 K pot.
R70	500 K 1W.
R71	500 K 1W.
R72	100 K 1W.
R73	680 K ½W.±5%
R74	250 K ½W.±5%
R75	Deleted
R76	1 meg. 1W.
R77	500 K 1W.
R78	1 K pot.

*2-75 K 1W. in parallel

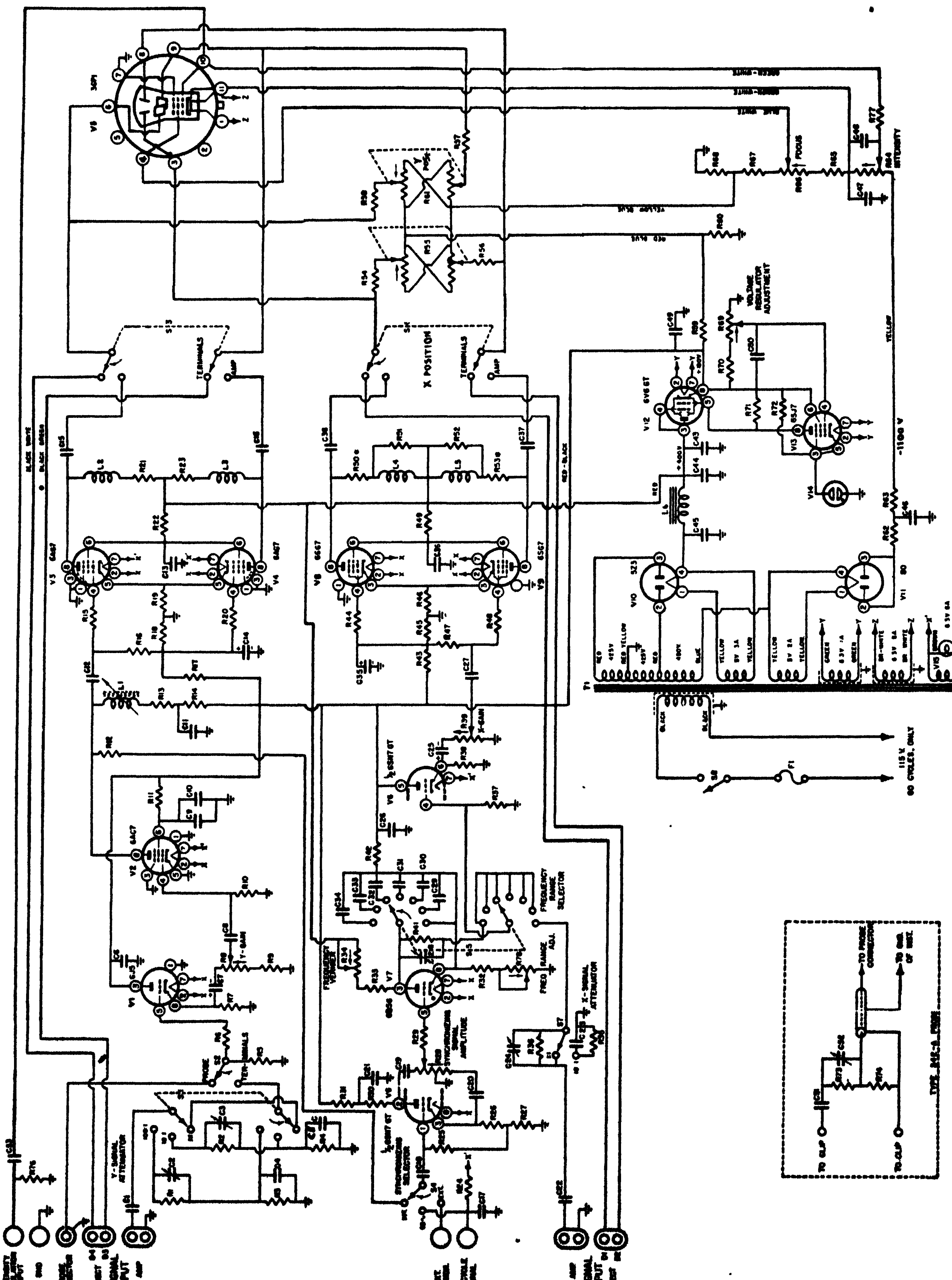


Fig. 22-10.—Schematic of Du Mont Model 224-A.

Courtesy Du Mont Labs.

DU MONT MODEL 241

FREQUENCY RESPONSE
Vertical Amplifier 20 cps to 2 Mc (uniform within 3 db)
Horizontal Amplifier 5 cps to 100 cps (uniform within 3 db)
Sweep Circuit 15 cps to 30,000 cps

DEFLECTION FACTORS
Vertical Amplifier Without probe 0.07 (with 0.7) rms volts/inch
Vertical-Deflection Plates 22 rms volts/inch
Horizontal Amplifier 0.7 rms volts/inch
Horizontal-Deflection Plates 22 rms volts/inch

LINE RATING 115 volts, 60 cps

TUBE COMPLEMENT

Type	*Function*
6J5 (V1)	Cathode Follower, Vertical Amplifier Input
6AC7 (V2)	Voltage Amplifier, Vertical Amplifier
6AG7 (V3, V4)	Push-Pull Vertical Driver Amplifier
5JP1 (V5)	Cathode-Ray Tube
6SN7GT (V6, V13)	Polarity Changer and Voltage Amplifier, Vertical Amplifier
6Q5G (V7)	Sweep Oscillator
6SN7GT (V8, V9)	Voltage Amplifier and Vertical Driver
6SG7 (V10)	Horizontal Amplifier Driver
6SG7 (V11)	Horizontal Amplifier Driver
6AC7 (V12)	Z-Axis Amplifier
6SN7GT (V14, V15)	Polarity Control, Z-Axis
5Z3 (V16)	Full-Wave Rectifier (Low Voltage)
80 (V17)	Half-Wave Rectifier (High Voltage)
6V6GT (V18)	Electronic Voltage Regulator
6SJ7 (V19)	Electronic Voltage Regulator

The schematic circuit diagram of Model 241 is shown in Fig. 22-11. This equipment is similar, in many respects, to the Du Mont Model 224-A previously shown. However, Model 241 does have some differences which will be discussed.

V12, V13, V14, and *V15* represent a Z-axis grid-control system for the cathode-ray tube *V5. V12* drives the cathode-ray-tube grid through *C42* and, in turn, receives its excitation from *R73.*

The setting of *R73* determines the polarity of the signal on the *V12* grid and across *R70.* If *R73* is set so that the voltage between pin 6 of *V15* and ground is applied to *C44* and *R70,* one polarity is obtained. If *R73* is set to obtain a portion or all of the voltage between pin 3 of *V14* and ground, a signal opposite in phase, or polarity, is obtained with reference to the polarity of the first signal (pin 6, *V15.*)

V14-V15, therefore, is in combination with *R73,* a phase-shift or polarity control for the Z-axis. The driving voltage for this stage is obtained from *V13* or direct from the input signal terminals marked Z-INPUT on the diagram. When *S7,* the Z-signal selector, is set to its middle position, *C45* is connected to the plate of *V13* which acts then as driver for *V14.*

V13 receives its grid excitation through *C43* and *C32* and is simply a grid (Z-axis) voltage amplifier. The grid of *V8* is driven by the 6Q5G sweep oscillator in the conventional manner or may receive its excitation through the X-axis signal input terminals and attenuator system consisting of *C17, C18, C19, R26,* and *S3.*

PARTS LIST FOR DU MONT MODEL 241

C1	0.5 μf. 600V.
C2	3-12 μμf.
C3	3-12 μμf.
C4	70 μμf. 500V.
C5	.001 μf. 500V.
C6	0.5 μf. 600V.
C7	100 μf. 50V.
C8	0.1 μf. 1000V.
C9	1 μf. 600V.
C10	0.5 μf. 200V.
C11	4 μf. 600V.
C12	25 μf. 50V.
C13	0.5 μf. 600V.
C14	0.1 μf. 1000V.
C15	0.1 μf. 1000V.
C16	.05 μf. 400V.
C17	0.5 μf. 600V.
C18	3-12 μμf.
C19	70 μμf. 500V.
C20	.05 μf. 400V.
C21	1 μf. 200V.
C22	0.1 μf. 1000V.
C23	0.1 μf. 1000V.
C24	3-12 μμf.
C25	150 μμf. 500V.
C26	600 μμf. 500V.
C27	.0025 μμf. 500V.
C28	.01 μf. 400V.
C29	.04 μf. 400V.
C30	0.15 μf. 400V.
C31	0.1 μf. 1000V.
C32	200 μμf. 500V.
C33	100 μf. 50V.
C34	0.5 μf. 600V.
C35	25 μf. 50V.
C36	0.5 μf. 600V.
C37	0.1 μf. 1000V.
C38	0.1 μf. 1000V.
C39	.05 μf. 400V.
C40	0.5 μf. 600V.
C41	0.5 μf. 600V.
C42	.05 μf. 1600V.
C43	.05 μf. 400V.
C44	0.1 μf. 1000V.
C45	0.1 μf. 1000V.
C46	0.1 μf. 1000V.
C47	4 μf. 600V.
C48	4 μf. 600V.
C49	4 μf. 600V.
C50	1 μf. 200V.
C51	0.5 μf. 1500V.
C52	0.5 μf. 1500V.
C53	0.5 μf. 1500V.
C54	1 μf. 400V.
C55	0.5 μf. 600V.
C56	.05 μf. 400V.
C57	3-12 μμf.
C58	0.25 μf. 400V.
F1	3 Amp. Fuse
L1	70-250 μh.
L2	130-500 μh.
L3	130-500 μh.
L4	7-19 mh.
L5	7-19 mh.
L6	33-100 μh.
L7	90-196 mh.
L8	19 h. 150 ma.
R1	2 meg. ½W. ±5%
R2	250 K. ½W. ±5%
R3	2 meg. ½W. ±5%
R4	20 K. ½W. ±5%
R5	1 K pot.
R6	1 K ½W.
R7	1 K. pot.
R8	110 ohms ½W.
R9	25 meg. 1W.
R10	100 K ½W.
R11	2 K. 1W.
R12	75 K. 1W.
R13	10 K. 3W.
R14	10 K. ½W.
R15	75 K. 1W.
R16	50 ohms ½W.
R17	50 K ½W.
R18	50 ohms ½W.
R19	470 ohms 3 W.
R20	15 K. 10W.
R21	4 K. 10W. non. ind.
R22	4 K. 10W. non. ind.
R23	500 K. ½W.
R24	5 meg. ½W.
R25	5 meg. ½W.
R26	2 meg. ½W. ±5%
R27	250 K. ½W. ±5%
R28	2 meg. ½W.
R29	50 ohms ½W.
R30	100 K ½W.
R31	500 K ½W.
R32	100 K 1W.
R33	10 K 1W.
R34	2 K ½W.
R35	8 K 1W.
R36	200 K pot. C. T.
R37	20 K ½W.
R38	3 meg. ½W.
R39	1 K 1W.
R40	82 K 3 W.
R41	2 meg. ½W.
R42	500 K 1W.
R43	4 meg. pot. ±10%
R44	2 meg. ½W.
R45	10 K 1W.
R46	1 K ½W.
R47	1 K ½W.
R48	10 K pot.
R49	1 K ½W.
R50	75 K 1W.
R51	10 K ½W.
R52	50 ohms ½W.
R53	2 meg. ½W.
R54	50 ohms ½W.
R55	25 K 1W.
R56	82 K 3 W.
R57	39 K 3 W.
R58	39 K 3 W.
R59	10 K ½W.
R60	10 K ½W.
R61	5 meg. ½W.
R62	500 K ½W.
R63	5 meg. ½W.
R64	2 K 1W.
R65	10 K 1W.
R66	75 K 1W.
R67	10 K ½W.
R68	2.5 K 1W.
R69	1 meg. ½W.
R70	25 meg. 1W.
R71	1 meg. ½W.
R72	1 K ½W.
R73	2 K pot. C.T.
R74	1 meg. ½W.
R75	200 K 1W.
R76	100 K 1W.
R77	4 meg. } dual pot.
R78	4 meg. } dual pot.
R79	2 K 1W.
R80	100 K 1W.
R81	100 K 1W.
R82	500 K pot.
R83	500 K 1W.
R84	100 K 1W.
R85	500 K 1W.
R86	250 K 1W.
R87	750 K 1W.
R88	500 K pot.
R89	100 K 1W.
R90	200 K pot.
R91	500 K 1W.
R92	1 meg. ½W. ±5%
R93	120 K. ½W. ±5%

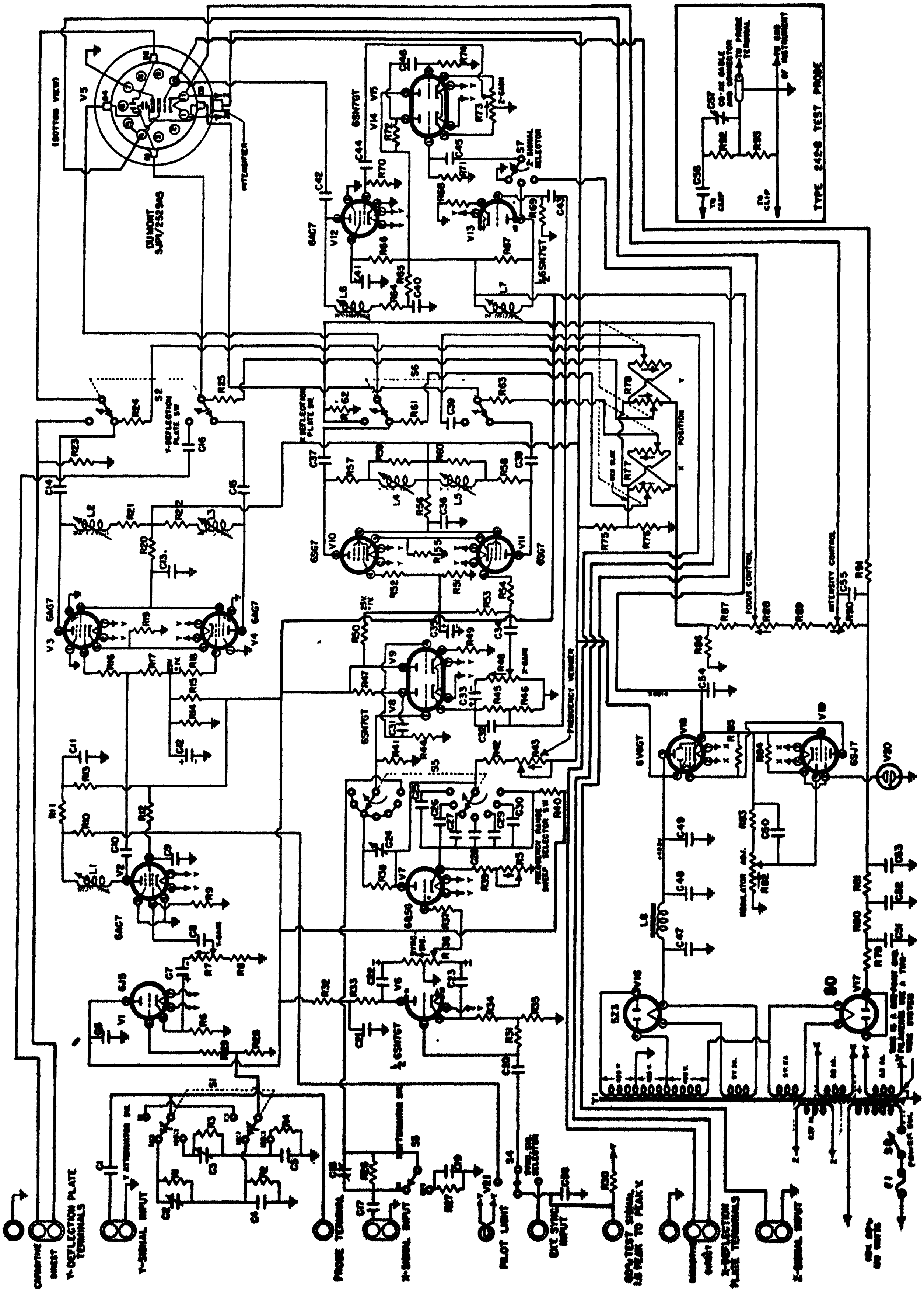


Fig. 22-11.—Schematic of Du Mont Model 241.

Courtesy Du Mont Labs.

DU MONT MODEL 247

Frequency Response
Vertical Amplifier Flat, 2 cps to 100 kc
Horizontal Amplifier Flat, 1 cycle to 30 kc
Sweep Circuit ½ cycle to 50 kc

Deflection Factors
Vertical Amplifier 0.05 rms volts/inch
Vertical-Deflection Plates 25 rms volts/inch
Horizontal Amplifier 0.5 rms volts/inch
Horizontal-Deflection Plates 21 rms volts/inch

Line Rating 115-230 volts, 40-60 cps

Tube Complement

Type	*Function*
6J5 (V1)	Input Stage Cathode Follower
6SH7 (V2)	Vertical Voltage Amplifier
6J5 (V3)	Cathode-Follower Driver
6AG7 (V4, V5)	Vertical Push-Pull Deflection Amplifier
5CP1 (V6)	Cathode-Ray Tube
6SN7GT (V7, V8)	Grid Signal Input Voltage Amplifier and Polarity Changer
6SN7GT (V9, V10)	Blanking Amplifier
6AG7 (V11)	Z-Axis Amplifier
6SN7GT (V12)	Blanking Control
6SN7GT (V13, V14)	Sync Amplifier
6SL7GT (V15, V16)	Diode Control and Polarity Changer
6Q5G (V17)	Sweep Oscillator
6SN7GT (V18, V21)	Cathode Follower and Vertical Amplifier Input
6SN7GT (V19, V20)	Diode Control and Blanking Amplifier
6SN7GT (V22, V23)	External Sync and Blanking Control Amplifier (Sweep Linearity Compensation)
6SL7GT (V24, V25)	Differential Amplifier
6AG7 (V26, V27)	Push-Pull Horizontal-Deflection Amplifier
VR90/30 (V28)	Neon Tube Voltage Regulator
VR150 (V29)	Voltage Regulator
6SL7GT (V30, V31)	Regulator Control Tube
6SL7GT (V32, V33)	Regulator Control
2X2 (V34, V35)	Half-Wave Rectifiers
6X5 (V36, V37)	Negative Rectifiers
5U4G (V38)	Positive Rectifier
6V6GT (V39, V40)	Series Voltage Regulator

The schematic circuit diagram of Model 247 is shown in Fig. 22-12.

Amplifier System

There is a direct d-c path from the plates of vertical amplifiers *V1*, *V2*, and *V3*, to the grid of *V4* which permits, in this section of the circuit, Y-axis or vertical position control of the spot on the screen of the cathode-ray tube by means of control *R19*. *C12*, a 50-μf capacitor, and *C13* (2,500 μμf) keep the impedance of the *R19* section of the circuit at a low value.

The horizontal-deflection amplifier is similar to the vertical amplifier. The d-c horizontal positioning is controlled by *R164*. Note that the *V26* grid is connected to the *V8* plate circuit and a source of positive potential through *R163*, *R164*, and *R166*, so that the *V26* grid may be made more or less positive as desired for positioning.

The Beam Control Circuit

A special circuit is used to vary the beam intensity in accordance with the sweep when used on single or repetitive sweep. Through its use, the cathode-ray-tube screen is made dark when the sweep is operating. Basically, this circuit uses a d-c amplifier which is connected to the cathode of the cathode-ray tube. The amplifier is driven by a pulse obtained from the sweep circuit which permits raising the beam intensity for the duration of the sweep cycle.

When switching from recurrent to single sweep, the voltage change cuts the cathode-ray-tube electron beam off. When the single sweep is begun, the beam returns to its former intensity. This is determined by the intensity-control setting and does not change during the sweep cycle. The intensity-control setting and the operation of the beam-control circuit are entirely independent.

Gas Triode (6Q5G) Operation

In the operation of this beam-control circuit, the gas triode *V17* is so biased that a higher plate potential is required to cause current conduction than is the case during recurrent operation. The 6Q5G is stopped from reaching the breakdown potential by the electronic control action of diode *V19*. The bias on this diode determines the actual potential on the 6Q5G plate.

The action of the gas triode is such that if its grid is driven by a positive pulse the bias on it is reduced and the tube becomes conductive. For pulses of shorter time duration than the time allowed for sweep, the tube returns to its former condition and is prepared for the succeeding cycle of its operation.

Current flows through the control diode *V19*, determined by *R98* and *R99*, during the time the 6Q5G is nonconductive. Conversely, no current flows through *V19* during the period the 6Q5G gas triode is conducting and during the time the sweep capacitor is charging up to the diode conducting potential. Consequently, a current flows through the resistance in series with the diode cathode which is virtually of constant value for a given control setting, except during the sweep cycle. At that time the cathode will be at its bias potential. The developed pulse controls the potential of the cathode-ray-tube cathode, and therefore, the cathode-ray-tube bias and intensity of the electronic light beam.

By varying the diode circuit resistance in proportion as the 6Q5G plate circuit resistance changes, the voltage across the resistance in the diode electronic control circuit is made substantially constant and is independent of the actual current flow.

Differential Amplifier

The tube 6SL7GT (*V24* and *V25*) is used as a differential amplifier to amplify only the potential differences across the diode load resistor (*R151* and *R152*) so that the setting of the diode bias voltage (single sweep adjustment) control *R153* does not affect the d-c operating levels of the following stages.

The output of the differential amplifier, amplified and reversed in phase from the input, is attenuated to zero level with respect to ground and fed directly to the grid of the following stage (*V20* one section of type 6SN7GT). The pulse level at this point is sufficient to operate the grid of *V20* between zero bias and cutoff, so that the plate potential of *V20* varies between very wide limits, from approximately 50 to 400 volts. A wide voltage variation is required for *V20*, since the signal must be attenuated up to the high negative potential at which the cathode of the cathode-ray tube operates, while still permitting the tube to swing from zero bias to cutoff.

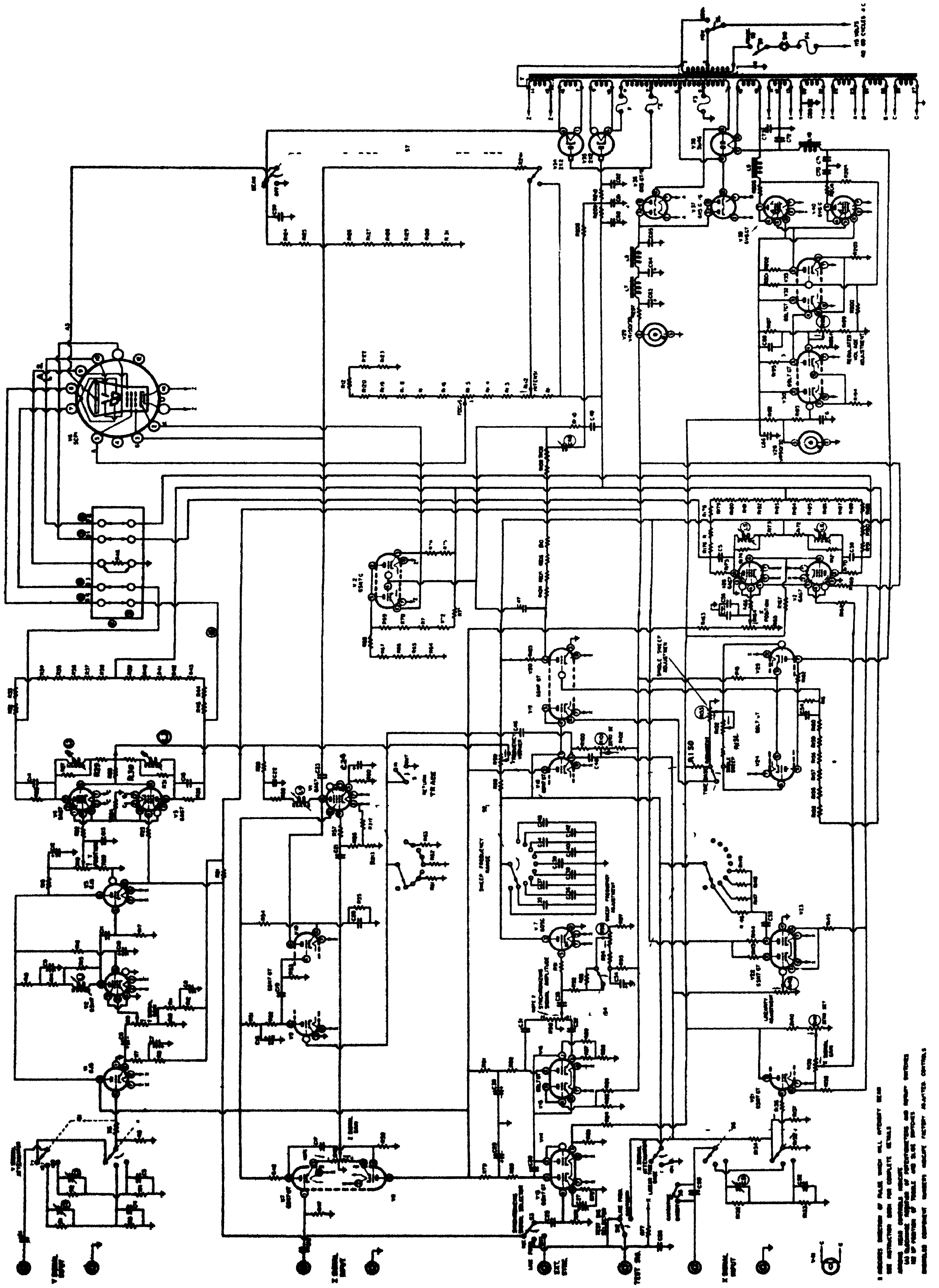
PARTS LIST FOR DU MONT MODEL 247

C1	0.5 µf.	600 V.
C2	3-12 µµf.	500 V.
C3	3-12 µµf.	500 V.
C4	100 µµf.	500 V.
C5	0.001 µf.	500 V.
C6	100 µf.	50 V. elec.
C7	24 µf.	350 V. elec.
C8	24 µf.	350 V. elec.
C9	16 µf.	450 V. elec.
C10	4 µf.	450 V. elec.
C11	0.5 µf.	600 V.
C12	50 µf.	25 V. elec.
C13	2200 µµf.	500 V.
C14	0.5 µf.	600 V.
C15	0.5 µf.	600 V.
C16	0.25 µf.	600 V.
C17	0.25 µf.	600 V.
C18	8 µf.	450 V. elec.
C19	0.25 µf.	600 V.
C20	100 µf.	50 V. elec.
C21	0.25 µf.	600 V.
C22	4 µf.	600 V.
C23	0.05 µf.	3000 V.
C24	100 µf.	50 V. elec.
C25	0.25 µf.	600 V.
C26	0.25 µf.	600 V.
C27	100 µf.	50 V. elec.
C28	0.05 µf.	400 V.
C29	8 µf.	450 V. elec.
C30	8 µf.	450 V. elec.
C31	0.25 µf.	600 V.
C32	0.25 µf.	600 V.
C33	0.25 µf.	600 V.
C34	50 µf.	25 V. elec.
C35	8 µf.	600 V.
C36	4 µf.	600 V.
C37	1 µf.	600 V.
C38	0.2 µf.	400 V.
C39	0.05 µf.	400 V.
C40	0.01 µf.	400 V.
C41	2200 µµf.	500 V.
C42	680 µµf.	500 V.
C43	220 µµf.	500 V.
C44	Omitted	
C45	47 µµf.	500 V.
C46	100 µµf.	500 V.
C47	47 µµf.	2500 V.
C48	60-240 µµf.	300 V.
C49	0.5 µf.	600 V.
C50	0.5 µf.	600 V.
C51	3-12 µµf.	500 V.
C52	100 µµf.	500 V.
C53	0.5 µf.	600 V.
C54	150 µµf.	500 V.
C55	50 µf.	25 V. elec.
C56	2200 µµf.	500 V.
C57	0.5 µf.	600 V.
C58	0.5 µf.	600 V.
C59	0.1 µf.	3000 V.
C60	½ µf.	2000 V.
C61	0.5 µf.	3000 V.
C62	0.5 µf.	3000 V.
C63	16 µf.	450 V. elec.
C64	16 µf.	450 V. elec.
C65	0.5 µf.	1500 V.
C66	24 µf.	350 V. elec.
C67	0.5 µf.	600 V.
C68	0.05 µf.	400 V.
C69	0.005 µf.	600 V.
C70	4 µf.	600 V.
C71	4 µf.	600 V.
C72	4 µf.	1000 V.
C73	4 µf.	1000 V.
F1	50 ma. fuse,	2500 V.
F2	50 ma. fuse,	2500 V.

F3	1 amp. fuse,	250 V.
F4	5 amp. fuse,	250 V.
L1	150-500 µh.	
L2	800-2500 µh.	
L3	800-2500 µh.	
L4	150-500 µh.	
L5	7½-19 mh.	±3%
L6	7½-19 mh.	±3%
L7	10 h. 50 ma.	dual
L8	10 h. 50 ma.	
L9	19 h. 150 ma.	
L10	19 h. 150 ma.	
R1	2 meg.	½W.±5%
R2	2 meg.	½W.±5%
R3	240 K	½W.±5%
R4	20 K	½W.±5%
R5	2.2 meg.	½W.
R6	47 ohm	½W.
R7	100 K	1W.
R8	22 K	1W.
R9	10 K pot.	
R10	1.1 K	½W.±5%
R11	27 K	1W.
R12	47 K	1W.
R13	15 K	1W.
R14	4.7 K	1W.
R15	10 K	½W.
R16	100 K	1W.
R17	2.2 meg.	½W.
R18	100 K	1W.
R19	5 K pot.	
R20	1.8 K	½W.
R21	100 K	1W.
R22	47 ohm	½W.
R23	47 ohm	½W.
R24	270 ohm	5W.NI±5%
R25	1 meg.	1W.
R26	1 meg.	1W.
R27	10 K	½W.
R28	30 K	10W.±5%
R29	10 K	20W.±5%
R30	8 K	20W.±5%
R31	10 K	½W.
R32	1 meg.	1W.
R33	1 meg.	1W.
R34	1 meg.	1W.
R35	1 meg.	1W.
R36	1 meg.	1W.
R37	1 meg.	1W.
R38	1 meg.	1W.
R39	1 meg.	1W.
R40	1 meg.	1W.
R41	1 meg.	1W.
R42	1 meg.	1W.
R43	1 meg.	1W.
R44	1 meg.	1W.
R45	1 meg.	1W.
R46	10 meg.	½W.
R47	1 meg.	½W.
R48	1 K	1W.
R49	2 K C.T. pot.	
R50	1 meg.	½W.
R51	22 K	1W.
R52	10 K	1W.
R53	1 meg.	½W.
R54	15K	10W.±5%
R55	2.2 K	1W.
R56	47 K	½W.
R57	47 ohm	½W.
R58	3 K	10W.NI±5%
R59	5 K	10 W.±5%
R60	180 ohm	1W.
R61	100 K	½W.
R62	47 K	½W.

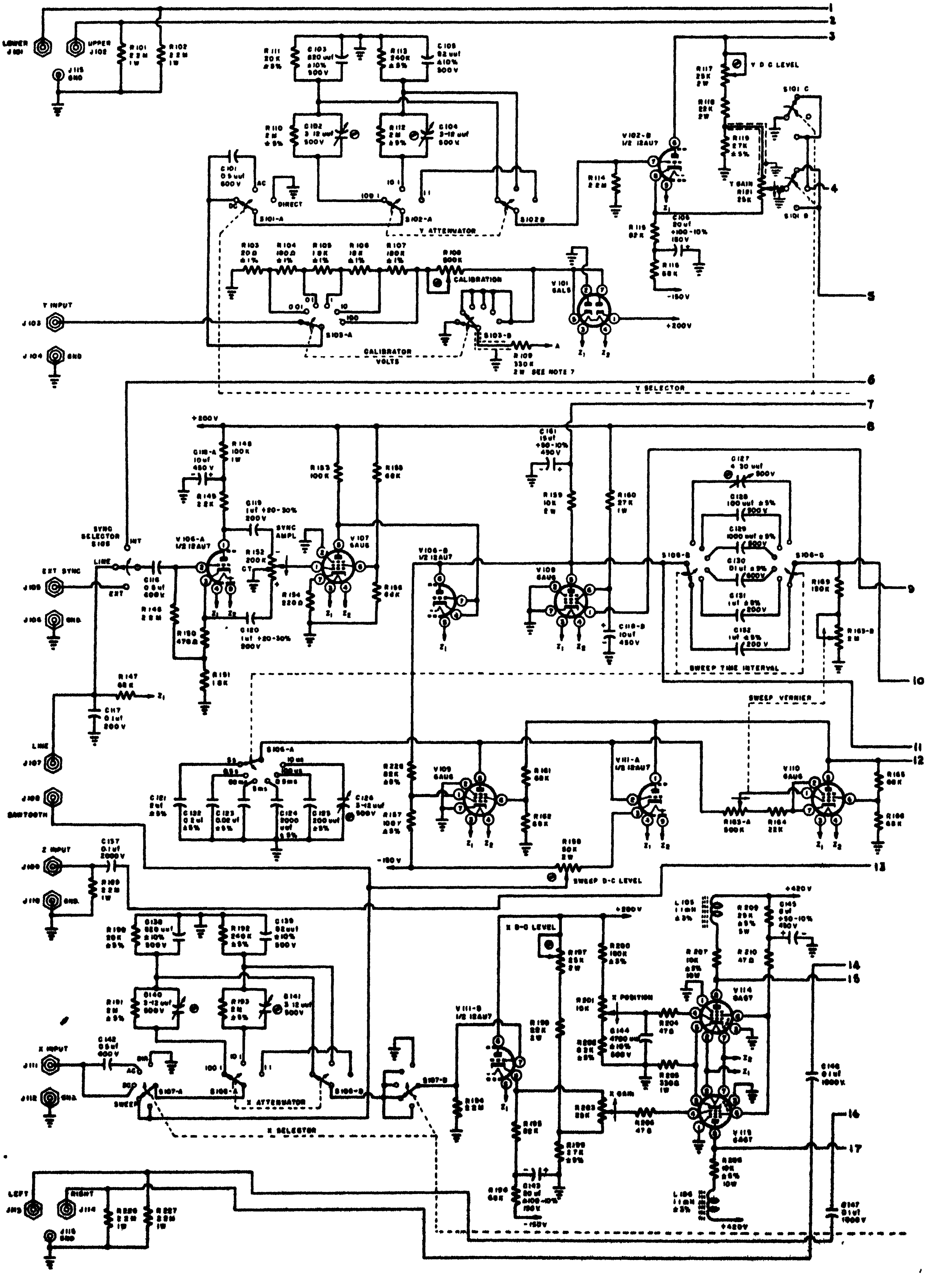
R63	10 K	½W.
R64	100 K	1W.
R65	100 K	1W.
R66	100 K	1W.
R67	100 K	1W.
R68	100 K	1W.
R69	100 K	1W.
R70	100 K	1W.
R71	100 K	1W.
R72	100 K	1W.
R73	100 K	1W.
R74	180 K	1W.
R75	180 K	1W.
R76	1 meg.	½W.
R77	15 K	½W.
R78	10 K	1W.
R79	22 K	1W.
R80	47 K	1W.
R81	10 K	1W.
R82	10 K	1W.
R83	1.8 K	½W.
R84	180 K	1W.
R85	1.8 K	½W.
R86	150 K	1W.
R87	1.8 K	½W.
R88	8.2 K	1W.
R89	2.2 meg.	½W.
R90	200 K C.T. Pot.	
R91	47 K	½W.
R92	100 K	½W.
R93	100 K	½W.
R94	470 ohm	½W.
R95	150 K	1W.
R96	2 K pot.	
R97	3.3 K	½W.
R98	470 K	1W.
R99	4 meg. pot.	dual
R152	100 K pot.	
R100	15 K	1W.
R101	10 K pot.	
R102	100 K	1W.
R103	100 K	3W.
R104	470 K	1W.
R105	470 K	1W.
R106	820 K	1W.
R107	820 K	1W.
R108	820 K	1W.
R109	820 K	1W.
R110	820 K	1W.
R111	47 K	1W.
R112	50 K pot.	
R113	47 K	1W.
R114	47 K	1W.
R115	100 K pot.	
R116	47 K	1W.
R117	47 K	1W.
R118	47 K	1W.
R119	47 K	1W.
R120	47 K	1W.
R121	47 K	1W.
R122	47 K	1W.
R123	47 K	1W.
R124	1 meg.	1W
R125	1 meg.	1W.
R126	1 meg.	1W.
R127	1 meg.	1W.
R128	1 meg.	1W.
R129	1 meg.	1W.
R130	1 meg	1W.
R131	1 meg.	1W.
R132	2 meg.	½W.±5%
R133	240 K	½W.±5%
R134	10 K	½W.
R135	2.2 meg.	½W.
R136	47 ohm	½W.
R137	10 K	½W.

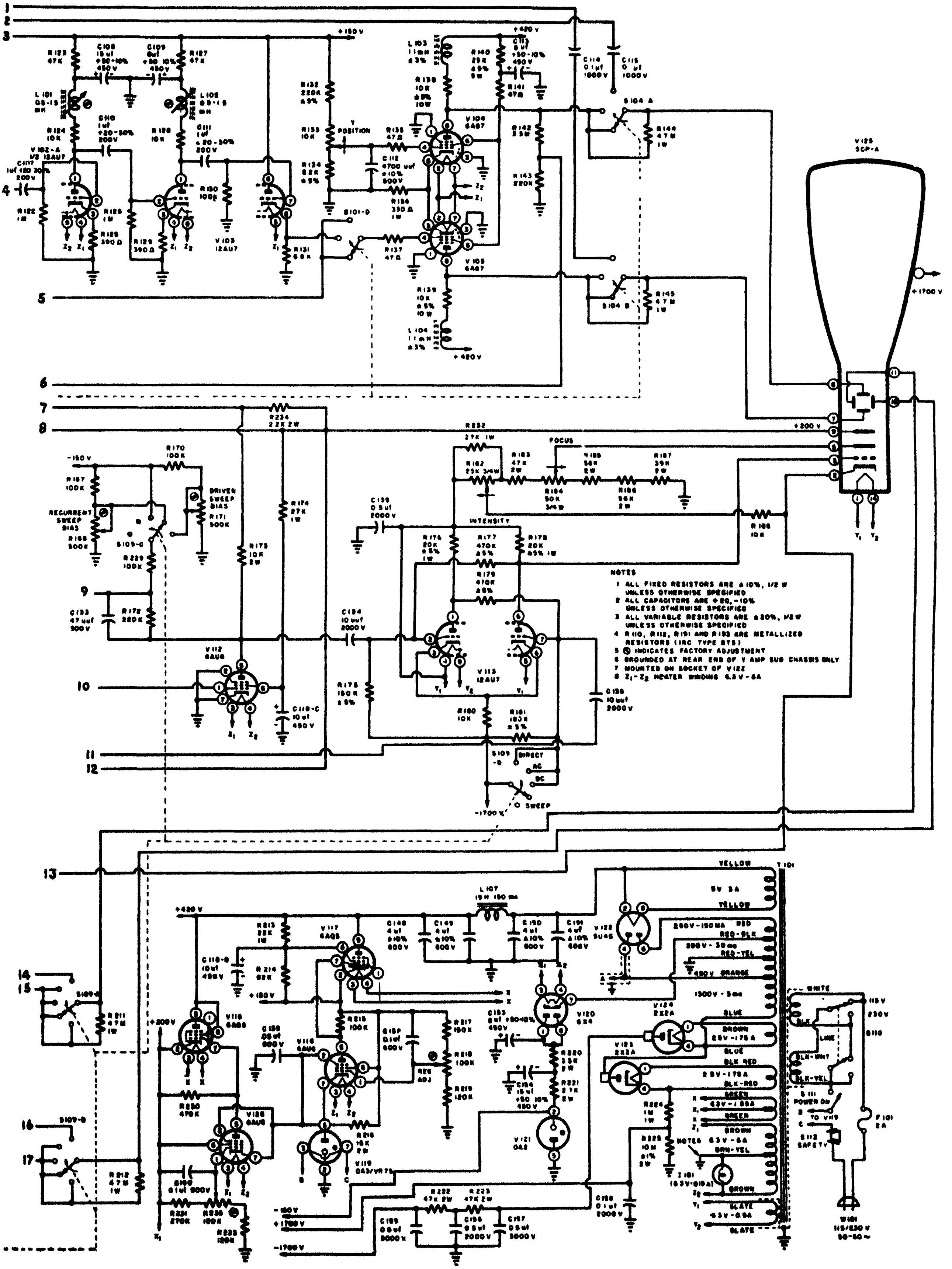
R138	100 K	1W.
R139	25 K pot.	
R140	22 K	3W.
R141	2 K pot.	
R142	500 K pot.	
R143	270 K	1W.
R144	270 K	1W
R145	150 K	1W.
R146	270 K	½W.
R147	470 K	½W.
R148	1 meg.	½W.
R149	2.2 meg.	½W.
R150	22 K	3W.
R151	12 K	1W.
R152	See R99	
R153	10 K pot.	
R154	100 K	1W.
R155	1 meg.	1W.
R156	1 meg.	1W.
R157	1 meg.	1W.
R158	1 meg.	1W.
R159	820 K	1W.
R160	470 K	1W.
R161	1 meg.	1W.
R162	100 K	1W.
R163	91 K	1W.±5%
R164	5 K pot.	
R165	1.5 K	½W.
R166	47 ohm	½W.
R167	39 K	3W.
R168	47 ohm	½W.
R169	10 K	10W.±5%
R170	1 meg.	1W.
R171	10 K	½W.
R172	30 K	10W.±5%
R173	30 K	10W.±5%
R174	10 K	½W.
R175	1 meg.	1W
R176	1 meg.	1W.
R177	1 meg.	1W
R178	1 meg.	1W.
R179	1 meg.	1W.
R180	1 meg.	1W.
R181	1 meg.	1W.
R182	1 meg.	1W.
R183	1 meg.	1W.
R184	1 meg.	1W.
R185	1 meg.	1W.
R186	1 meg.	1W.
R187	1 meg.	1W.
R188	1 meg.	1W.
R189	1 meg.	1W.
R190	1 meg.	1W.
R191	1 meg.	1W.
R192	10 K	3W.
R193	2.2 meg.	½W.
R194	82 K	1W.
R195	470 K	1W.
R196	100 K	1W.
R197	100 K	1W.
R198	500 K pot.	
R199	100 K	1W.
R200	47 K	1W.
R201	27 K	1W.
R202	470 K	1W.
R203	150 K	1W.
R204	47 ohm	½W.
R205	1 meg.	1W.
R206	47 ohm	½W.
R207	5 K	25W.±5%
R208	100 K	1W.
R209	39 K	3W.
R210	39 K	3W.
R211	1 meg.	½W.
R212	2 K	½ W.±5%
R213	100 ohm	½W.±5%



Courtesy Du Mont Labs

Fig 22-12 —Schematic of Du Mont Model 247.


LOWER J101
UPPER J102
J115 GND
R101 2.2 M 1W
R102 2.2 M 1W
Y D C LEVEL
S101 C
V102-B 1/2 12AU7
Y GAIN R121 25K
S101 B
Y ATTENUATOR
S101-A
S102-A
S102B
DIRECT
AC
DC
CALIBRATION
V101 OA2
Y INPUT J103
J104 GND
S103-A
S103-B
CALIBRATOR VOLTS
R109 330K 2W SEE NOTE 7
Y SELECTOR
+200V
-150V
SYNC SELECTOR S105
INT
LINE
EXT
EXT SYNC J105
J106 GND
V106-A 1/2 12AU7
SYNC AMPL
V107 6AU6
V106-B 1/2 12AU7
V108 6AU6
S106-B
S106-C
SWEEP TIME INTERVAL
SWEEP VERNIER
R147 68K
C117 0.1uf 200V
LINE J107
J108 SAWTOOTH
S106-A
V109 6AU6
V111-A 1/2 12AU7
V110 6AU6
SWEEP D-C LEVEL
Z INPUT J109
J110 GND
C137 0.1uf 2000V
X D-C LEVEL
X POSITION
X GAIN
V111-B 1/2 12AU7
V114 6AU7
V113 6AU7
+420V
X INPUT J111
J112 GND
S107-A
S108-A
S108-B
S107-B
X ATTENUATOR
X SELECTOR
SWEEP
LEFT J113
RIGHT J114
J115 GND
R226 2.2M 1W
R227 2.2M 1W
1
2
3
4
5
6
7
8
9
10
11
12
13
14
15
16
17



NOTES
1 ALL FIXED RESISTORS ARE ±10%, 1/2 W UNLESS OTHERWISE SPECIFIED
2 ALL CAPACITORS ARE +20, −10% UNLESS OTHERWISE SPECIFIED
3 ALL VARIABLE RESISTORS ARE ±20%, 1/2 W UNLESS OTHERWISE SPECIFIED
4 R 110, R 112, R 191 AND R 193 ARE METALLIZED RESISTORS (IRC TYPE BTS)
5 Ⓢ INDICATES FACTORY ADJUSTMENT
6 GROUNDED AT REAR END OF Y AMP SUB CHASSIS ONLY
7 MOUNTED ON SOCKET OF V 122
8 Z_1-Z_2 HEATER WINDING 6.3 V - 6A

Fig. 22-13.—Schematic of Du Mont Model 250-A.

Courtesy Du Mont Labs.

DU MONT MODELS 250, 250-A, 250-AH, 250-H

SPECIFICATIONS FOR MODEL 250-A

FREQUENCY RESPONSE

Vertical Amplifier (a.c.) 5 cps to 200 kc, within 10%
Vertical Amplifier (d.c.) 0 cps to 200 kc, within 10%
Horizontal Amplifier Identical to above

SWEEP CIRCUIT	MODELS 250 AND 250-H	MODELS 250-A AND 250-AH
Recurrent	1 to 150,000 cps (1 second to 6.7 microseconds)	5 seconds to 10 microseconds
Driven	1 second to 20 microseconds	5 seconds to 10 microseconds

DEFLECTION FACTORS

Vertical Amplifier (a.c.) 15 rms millivolts/inch
Vertical Amplifier with probe 150 rms millivolts/inch
Vertical Plates Direct 21 rms millivolts/inch
Vertical Amplifier (d.c.) 0.9 d-c millivolts/inch
Horizontal Amplifier (a.c.) 0.4 rms millivolts/inch
Horizontal Plates Direct 23 rms millivolts/inch
Horizontal Amplifier (d.c.) 1.1 d-c millivolts/inch

LINE RATING 115-230 volts, 50-60 cps

TUBE COMPLEMENT FOR MODELS 250-A, 250-AH

Type	*Function*
6AL5 (V101)	Voltage Calibrator
12AU7 (V102)	Vertical Amplifier
12AU7 (V103)	Vertical Amplifier
6AG7 (V104)	Vertical Deflection
6AG7 (V105)	Vertical Deflection
12AU7 (V106)	Sync Input
6AU6 (V107)	Sync Amplifier
6AU6 (V108)	Negative Gate
6AU6 (V109)	Switching Tube
6AU6 (V110)	Sweep Charging Tube
12AU7 (V111)	X Input, Sweep Output
6AU6 (V112)	Negative Signal Amplifier
12AU7 (V113)	Flip-Flop Oscillator
6AG7 (V114)	Horizontal Deflection
6AG7 (V115)	Horizontal Deflection
6AQ5 (V116)	Voltage Regulator
6AQ5 (V117)	Voltage Regulator
6AU6 (V118)	Voltage Regulator
OA3/UR75 (V119)	Voltage Regulator
6X4 (V120)	Low-Voltage Rectifier
OA2 (V121)	Voltage Regulator
5U4G (V122)	Full-Wave Rectifier
2X2A (V123)	Positive High-Voltage Rectifier
2X2A (V124)	Negative High-Voltage Rectifier
5CP-A (V125)	Cathode-Ray Tube
6AU6 (V126)	Voltage Regulator

The following discussion on Models 250-A and 250-AH also applies to early Models 250 and 250-H, respectively. The earlier models have a slightly different tube line-up, but are electrically similar to the latter, with the exception of the modified sweep characteristics, noted in the specifications. A schematic circuit diagram for Model 250-A, which applies with minor modifications to all four models, is shown in Fig. 22-13.

The Model 250-AH (250-H) is identical to the Model 250-A (250) in outward appearance. However, at the back of the cabinet a connector is provided for the external intensifier supply. With a simple jumper arrangement inside the Model 250-AH (250-H) the intensifier of the type 5RP-A cathode-ray tube may be connected from an internal high-voltage supply instead of an external supply. The over-all accelerating potential is then 3,000 volts, and the instrument becomes electrically equivalent to the Model 250-A (250). Low-voltage operation is often desirable for applications where a low deflection factor is more important than high light output from the screen.

The 250-AH (250-H) uses a 5RP-A cathode-ray tube and an accelerating potential of 19,000 volts. The 250-AH (250-H) amplifier (maximum gain) deflection sensitivity is 25 rms millivolts per inch. With probe, it is 250 rms millivolts per inch. The d-c amplifier is rated at 1.7 d-c volts per inch and connection directly to the deflection plates results in 45 rms volts per inch deflection sensitivity.

The 250-AH (250-H) horizontal-deflection factors are as follows:

A-C Amplifier (Maximum Gain) 0.7 rms volts/inch
D-C Amplifier (Maximum Gain) 2 d-c volts/inch
Direct to Deflection Plates 48 rms volts/inch

The 250-A (250) is designed for general-purpose laboratory work. It has both a-c and d-c signal amplifiers. A built-in voltage calibrator allows accurate measurement of input signal amplitudes. The time-base generator furnishes a wide range of linear sweep speeds, both recurrent and driven. The intensity of the cathode-ray beam may be modulated from an external source to produce timing marks or reference points on the trace. The built-in circuit, which automatically controls the beam intensity, greatly simplifies the techniques in making photographic recordings.

Driven (Single) Sweeps

When the X-selector switch *S107* is in the driven-sweep position, a high negative bias potential is applied to the grid of *V108*, thus making the sweep circuit inoperative. The trigger voltage is received by the sync inverter *V106A* so that with the sync amplitude control *R152*, either the same or the opposite polarity of the triggering voltage can be selected for triggering the sweep.

Since a negative triggering voltage of relatively high amplitude is required by the gate generator, *V107* is used to amplify the trigger signal. Therefore, to trigger the circuit, a positive trigger signal is taken from *R152*, amplified, inverted, and fed through a diode *V106B* to the plate of *V108*. This negative signal is coupled to *V112* through the network *C127* to *C132* and *R163B*. This is amplified by *V112* and fed back regeneratively to *V108* to the point where *V108* is conducting and *V112* is cut off. This condition remains until the capacitors *C127* to *C132* discharge sufficiently through *R169* to permit *V112* to conduct again. This occurs in a second regenerative cycle as described above.

A negative trigger voltage is obtained from the plate of *V108* by differentiation of the gate signal. This trigger is fed to pin 7 of *V113*, causing that tube to cut off and the other half of the Eccles-Jordan circuit to conduct. At the same time it permits operation of the cathode-ray-tube grid, allowing the tube to be brightened to that level set by the intensity control. At the end of the gate, a second negative trigger is derived from the plate of *V112* by differentiation of the gate voltage and is fed to pin 2 of *V113*. This causes the Eccles-Jordan circuit to reverse, cutting off the cathode-ray tube and the spot on the screen.

Recurrent Sweeps

The recurrent-sweep operation is similar to the driven-sweep operation except for bias setting. When the gate generator (V108-V112) has recovered, *V108* will be held off until the capacitor *C133* reaches a voltage which allows *V108* to again start conducting, at which time it will repeat the sweep cycle. There is no "return trace blanking." The beam is always off except when it is sweeping from left to right.

DU MONT MODEL 256-D

FREQUENCY RESPONSE

Vertical Amplifier Down 3 db at 8 Mc

SWEEP CIRCUITS

A sweeps: 1,000, 100, 25, 10, 4 microseconds; in addition, a 4,500-microsecond sweep for observing the entire duty cycle at repetition rates above 300 per second.

R sweeps: 25, 10, 4 microseconds which may be delayed to cover any portion of the 100-microsecond *A* Sweep. The delay is read directly on a dial with an accuracy of ±0.1% in the 4-to-100-microsecond or 5-to-1,000-microsecond regions.

DEFLECTION FACTORS

Vertical Amplifier 0.25 rms volts/inch
Vertical-Deflection Plates 79 d-c volts/inch

LINE RATING 115 volts, 60 cps, ±10%

The schematic circuit diagram and tube complement of Model 256-D is shown in Fig. 22-14A. A block diagram is shown in Fig. 22-14B. This instrument, a relatively specialized apparatus, is a revised form of the 256-B *A/R* range scope. The time scale of the instrument has been changed from yards to microseconds.

Trigger Circuits

Triggering may be obtained either from an internal trigger generator of variable rate or from an external signal of positive or negative polarity. Timing markers are available for use only when the internal trigger generator is used. At this time, the markers appearing on the sweep are also available at the Z IN-MARKER OUT terminal.

Sweep-Delay Circuit, Pick-Off-Diode, and Delay Amplifiers

The sweep-delay circuit consists essentially of a compensated linear-sawtooth circuit which is connected to a *pick-off diode*. The voltage at which this diode conducts is controlled through a bleeder network by a ten-turn, helically wound potentiometer. A dial attached to this potentiometer permits reading off the delay time directly. The output of the pick-off diode is amplified by tubes *V105A* and *V105B* and serves as the delayed trigger for the rest of the circuits.

R-Gate and Sweep Generators

The circuit for the *R*-gate generator consists of a special blocking oscillator which provides a positive signal for intensifying the cathode-ray tube during the *R* sweeps, a negative signal to act as the movable marker, and a negative gate for initiating the *R* sweeps.

Crystal Oscillator

The crystal oscillator is a conventional crystal-controlled triode oscillator. The crystal frequency is 100 kc which gives timing marks of 10 μsec apart. It drives the 10-μsec marker generator.

Marker Generators

The marker generators are conventional blocking oscillators. The 10-μsec marker generator is triggered by the crystal oscillator. The 50-μsec marker generator is triggered by the 10-μsec markers. The output of the marker generators when used on the sweeps appear as intensity markers (—) or blanking markers (+), because they are impressed upon the *cathode* of the cathode-ray tube. The 10-μsec marker generator is driven by the pulses from the crystal oscillator. The 50-μsec marker generator *V110A* is a conventional blocking oscillator which is triggered from the 10-μsec marker generator by the positive 10-μsec markers impressed upon the grid of *V110A*.

Trigger Circuits

The *trigger-divider gate generator V110B* is a blocking oscillator triggered in the cathode circuit from the +50-μsec markers. This blocking oscillator divides the repetition rate and generates a 16-μsec gate. The output of the oscillator is a gate which is fed to the *trigger generator*.

One triode of *V111* is used as coincidence circuit for the trigger-divider gate and a 10-μsec marker. When a marker is coincident with the gate, the trigger generator is fired. Another triode, the trigger generator is a conventional blocking oscillator, but it is biased off to prevent free running. A negative trigger is used to feed the *A* gate, and a positive trigger is used to feed the trigger cathode-follower output tube. The *trigger output* is a conventional phase splitter and is used to furnish either a positive or negative trigger for external use. The positive trigger is also used to trigger the *R* gate and hence the *R* sweeps when an undelayed sweep is desired.

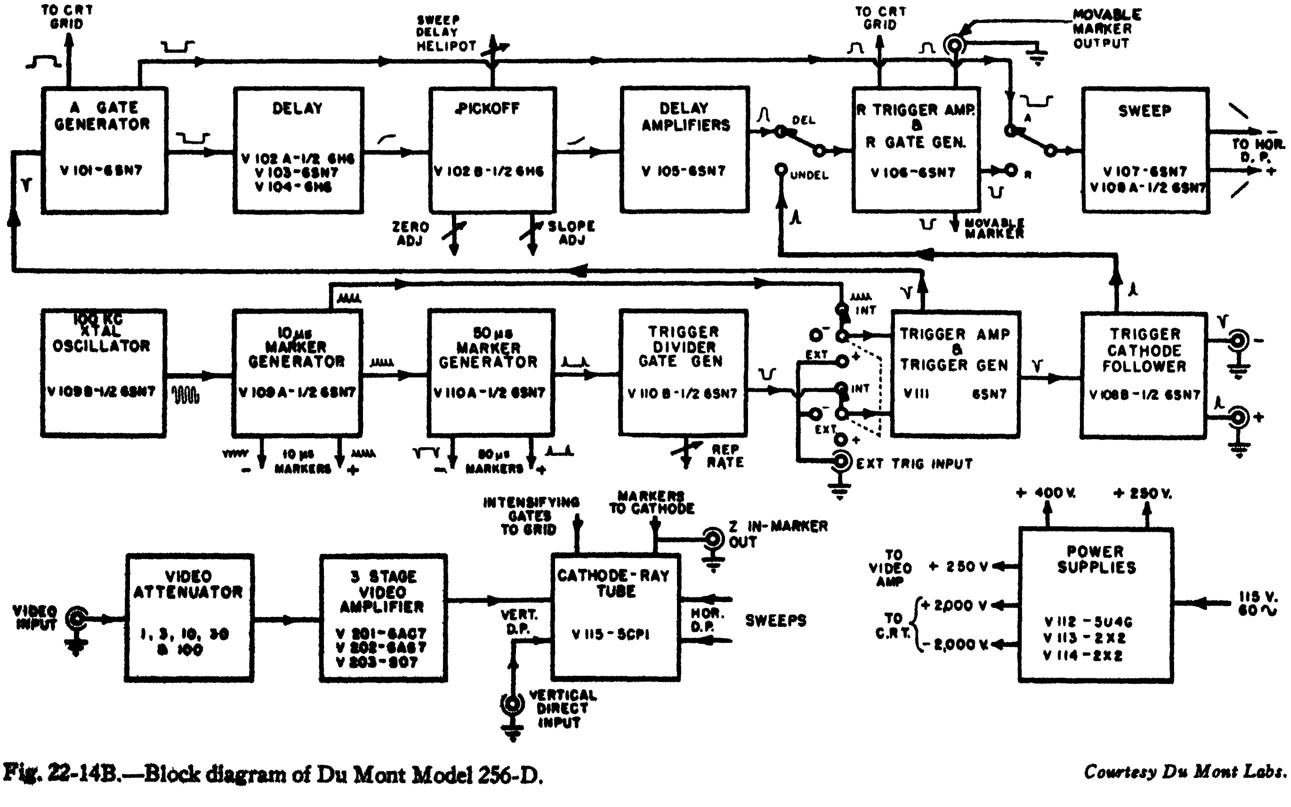
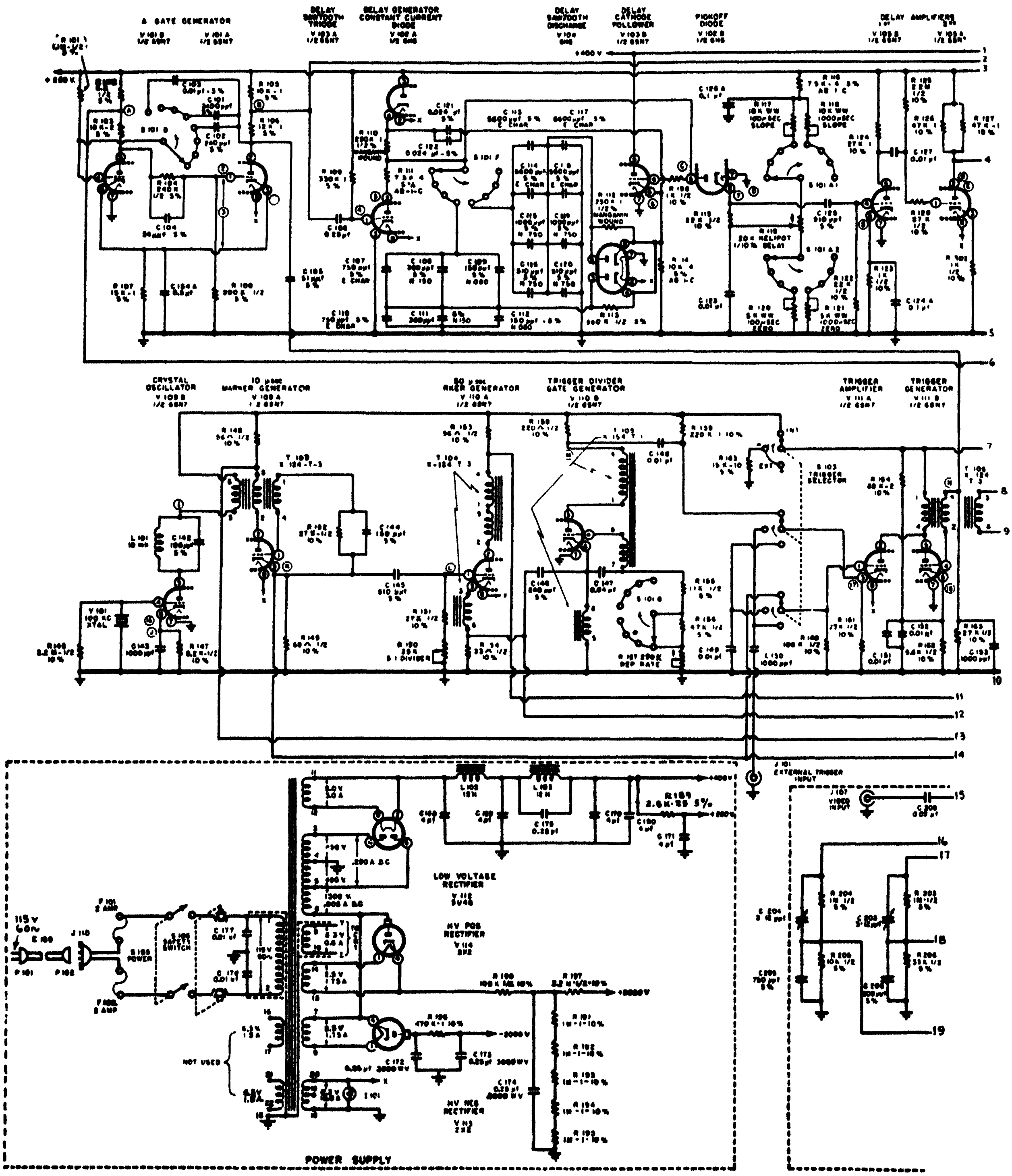


Fig. 22-14B.—Block diagram of Du Mont Model 256-D.

Courtesy Du Mont Labs.


A GATE GENERATOR
V 101 B 1/2 6SN7
V 101 A 1/2 6SN7
DELAY SAWTOOTH TRIODE
V 103 A 1/2 6SN7
DELAY GENERATOR CONSTANT CURRENT DIODE
V 102 A 1/2 6H6
DELAY SAWTOOTH DISCHARGE
V 104 6H6
DELAY CATHODE FOLLOWER
V 103 B 1/2 6SN7
PICKOFF DIODE
V 102 B 1/2 6H6
DELAY AMPLIFIERS
1st
2nd
V 105 B 1/2 6SN7
V 105 A 1/2 6SN7
+400 V
+250 V
CRYSTAL OSCILLATOR
V 109 B 1/2 6SN7
10 μsec MARKER GENERATOR
V 109 A 1/2 6SN7
50 μsec MARKER GENERATOR
V 110 A 1/2 6SN7
TRIGGER DIVIDER GATE GENERATOR
V 110 B 1/2 6SN7
TRIGGER AMPLIFIER
V 111 A 1/2 6SN7
TRIGGER GENERATOR
V 111 B 1/2 6SN7
S 103 TRIGGER SELECTOR
J 101 EXTERNAL TRIGGER INPUT
J 107 VIDEO INPUT
115 V 60~
LOW VOLTAGE RECTIFIER
V 112 5U4G
HV POS RECTIFIER
V 114 2X2
HV NEG RECTIFIER
V 113 2X2
R189 2.6K-25 5%
+2000 V
-2000 V
NOT USED
POWER SUPPLY
1
2
3
4
5
6
7
8
9
10
11
12
13
14
15
16
17
18
19

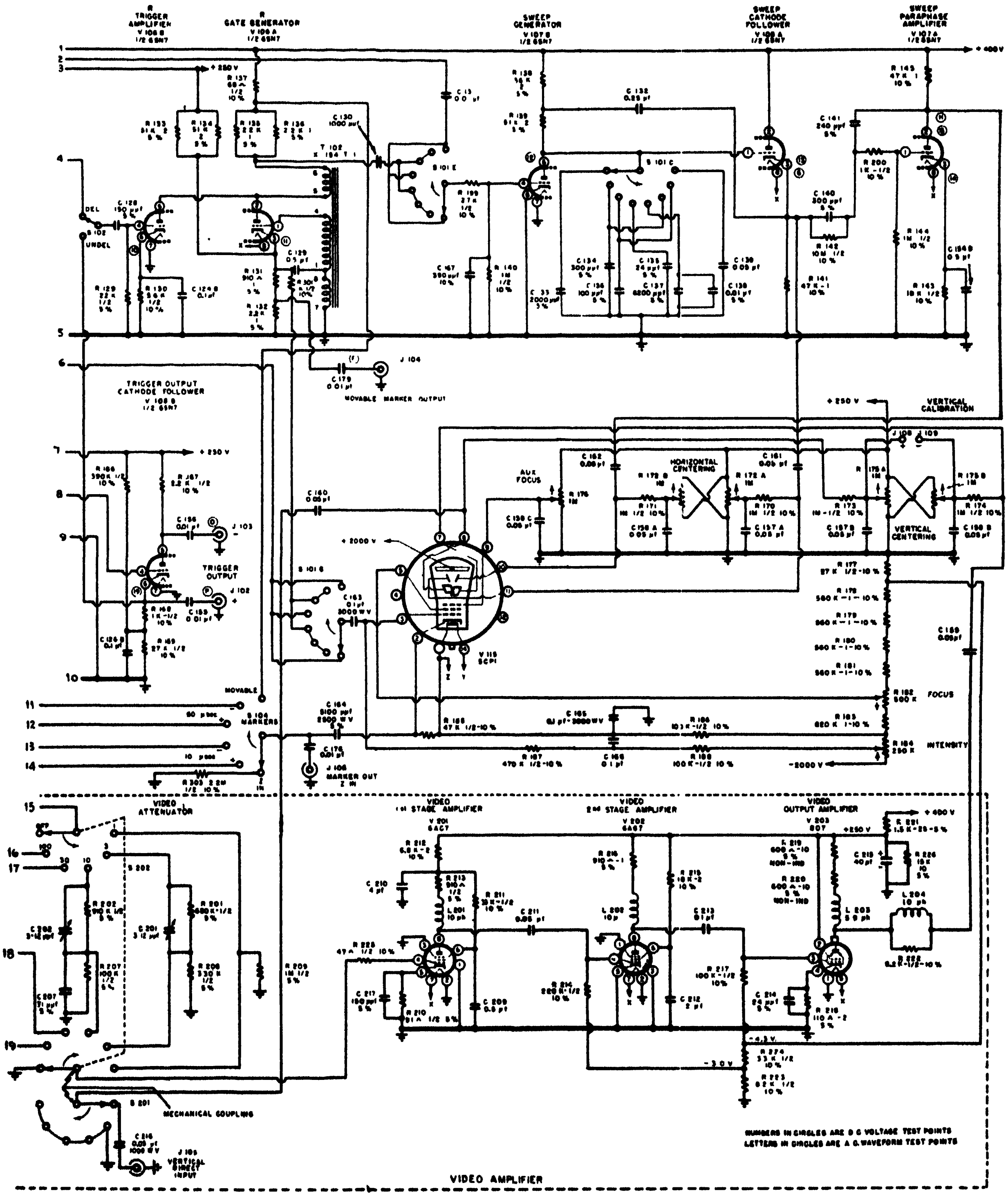

TRIGGER AMPLIFIER V 106 B 1/2 6SN7
GATE GENERATOR V 106 A 1/2 6SN7
SWEEP GENERATOR V 107 B 1/2 6SN7
SWEEP CATHODE FOLLOWER V 108 A 1/2 6SN7
SWEEP PARAPHASE AMPLIFIER V 107 A 1/2 6SN7
+400 V
+250 V
DEL
UNDEL
S 102
S 101 E
S 101 C
T 102
MOVABLE MARKER OUTPUT
J 104
TRIGGER OUTPUT CATHODE FOLLOWER V 108 B 1/2 6SN7
TRIGGER OUTPUT
J 103
J 102
VERTICAL CALIBRATION
HORIZONTAL CENTERING
VERTICAL CENTERING
AUX FOCUS
+2000 V
S 101 B
V 115 5CP1
FOCUS
INTENSITY
-2000 V
MOVABLE
MARKERS
S 104
J 106 MARKER OUT
VIDEO ATTENUATOR
S 202
S 201
MECHANICAL COUPLING
J 105 VERTICAL DIRECT INPUT
VIDEO 1st STAGE AMPLIFIER V 201 6AC7
VIDEO 2nd STAGE AMPLIFIER V 202 6AC7
VIDEO OUTPUT AMPLIFIER V 203 807
-4.3 V
-3.0 V
NUMBERS IN CIRCLES ARE D C VOLTAGE TEST POINTS
LETTERS IN CIRCLES ARE A C WAVEFORM TEST POINTS
VIDEO AMPLIFIER

Fig. 22-14A.—Schematic of Du Mont Model 256-D.

Courtesy Du Mont Labs.

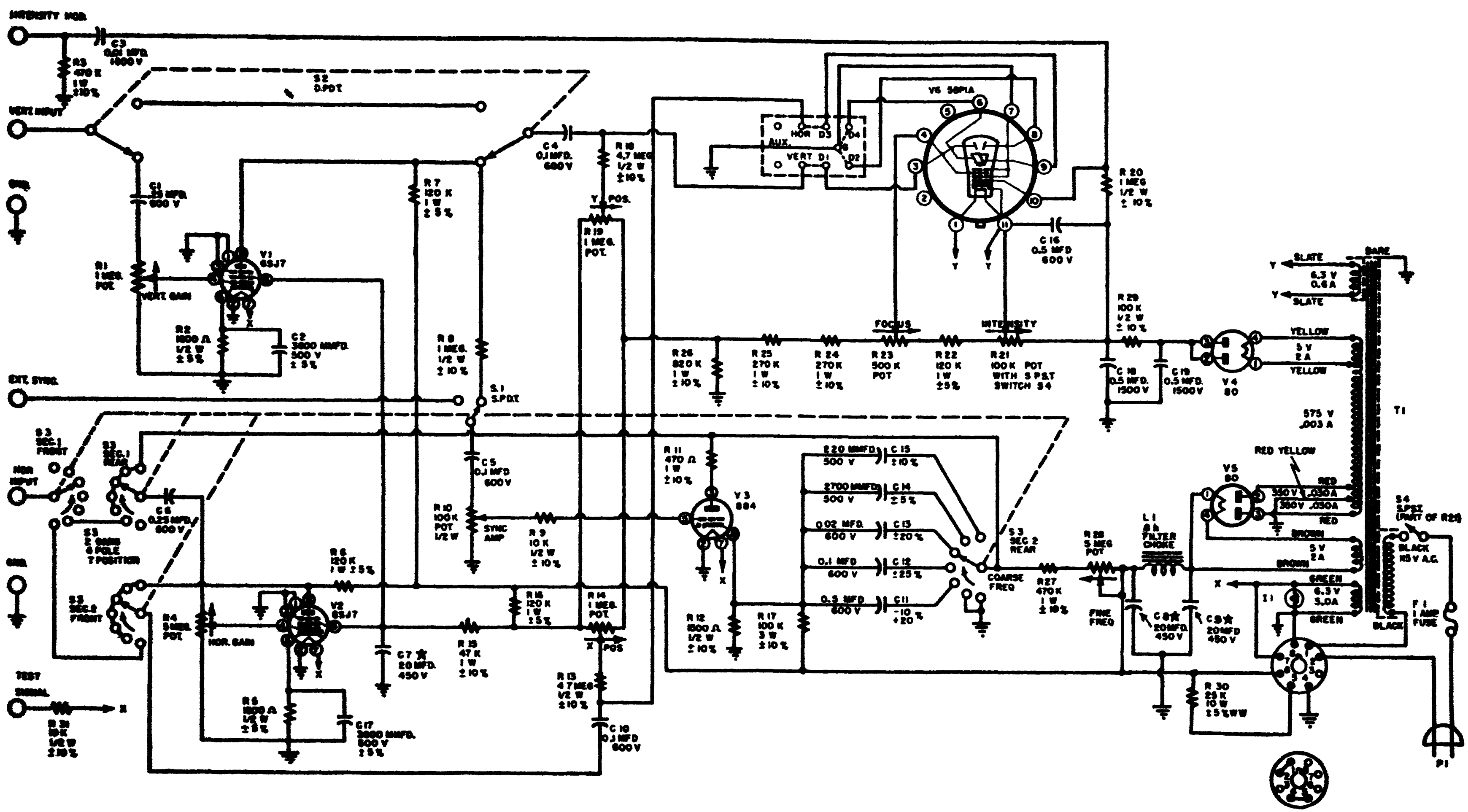


Fig. 22-15.—Schematic of Du Mont Model 274.

Courtesy Du Mont Labs.

DU MONT MODEL 274

Frequency Response
- Vertical Amplifier 20 cps to 50 kc, within ±20%
- Horizontal Amplifier 20 cps to 50 kc, within ±20%
- Sweep Circuit 8 cps to 30 kc

Deflection Factors
- Vertical Amplifier 0.65 rms volts/inch
- Vertical-Deflection Plates 18 rms volts/inch, ±20%
- Horizontal Amplifier 0.65 rms volts/inch
- Horizontal-Deflection Plates 18 rms volts/inch, ±20%

Line Rating 115 volts, 50-60 cps

Tube Complement

Type	*Function*
6SJ7 (V1)	Vertical Amplifier
6SJ7 (V2)	Horizontal Amplifier
884 (V3)	Sweep Oscillator
80 (V4)	Half-Wave Rectifier
80 (V5)	Full-Wave Rectifier
5BP (V6)	Cathode-Ray Tube

The schematic circuit diagram of this conventional unit is shown in Fig. 22-15.

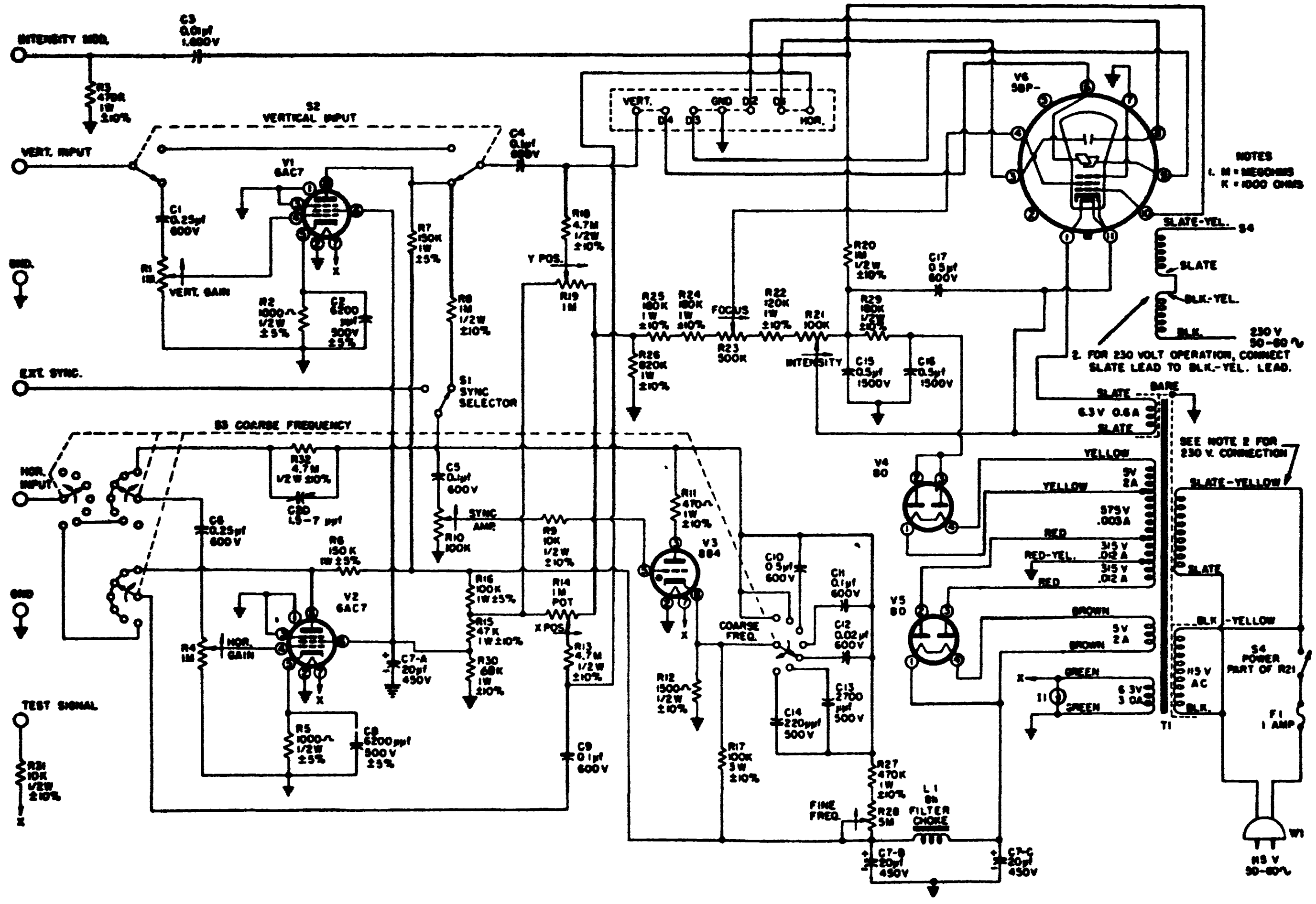


Fig. 22-16.—Schematic of Du Mont Model 274-A.

Courtesy Du Mont Labs.

DU MONT MODEL 274-A

Frequency Response
- Vertical Amplifier 20 cps to 200 kc
- Horizontal Amplifier 40 cps to 200 kc
- Sweep Circuit 8 cps to 30 kc

Deflection Factors
- Vertical Amplifier 0.2 rms volts/inch (max)
- Vertical-Deflection Plates 16 rms volts/inch
- Horizontal Amplifier 0.25 rms volts/inch
- Horizontal-Deflection Plates 18 rms volts/inch

Line Rating 115-230 volts, 50-60 cps

Tube Complement

Type	*Function*
6AC7 (V1)	Vertical Amplifier
6AC7 (V2)	Horizontal Amplifier
884 (V3)	Time-Base Generator
80 (V4)	High-Voltage Rectifier
80 (V5)	Low-Voltage Rectifier
5BP (V6)	Cathode-Ray Tube

The Model 274-A is similar to Model 274. However, its frequency response range is greater and gain higher. The tube line-up has also been modernized. A schematic circuit diagram of this conventional unit is shown in Fig. 22-16.

DU MONT MODEL 275-A

Frequency Response 2 cps to 30,000 cps
Circular Sweep Voltages are furnished from the external two-phase generator provided
Circle Diameter Adjustable from 0 to 4½ inches

Deflection Factor Less than 0.4 volt rms produces deflection to center of circle

Line Rating 115-230 volts, 50-60 cps

Tube Complement

Type	*Function*
6SL7-GT (V1)	Vertical Input and Phase Inverter
6SJ7 (V2)	Vertical Output
6SJ7 (V3)	Vertical Output
6SL7-GT (V4)	Horizontal Input and Phase Inverter
6SJ7 (V5)	Horizontal Output
6SJ7 (V6)	Horizontal Output
6SN7-GT (V7)	Radial Amplifier
6SN7-GT (V8)	Radial Amplifier and Cathode Follower
5CP1 (V9)	Cathode-Ray Tube
2X2-A (V10)	High-Voltage Rectifier
2X2-A (V11)	High-Voltage Rectifier
5Y3-GT (V12)	Low-Voltage Rectifier
6X5-GT (V13)	Negative-Voltage Rectifier
OC3 (V14)	Voltage Regulator
OC3 (V15)	Voltage Regulator

The schematic circuit diagram of Model 275-A is shown in Fig. 22-17. This instrument is of special design and construction designed primarily for the study of the various characteristics of rotating machinery. The time base in this instrument is circular and is produced in synchronism with the rotation of the machine being studied. The characteristic of the machine to be examined may be transformed into an electrical signal by means of a suitable device, and be applied to the input terminals. Its salient features will be plotted on the screen, and the relative position in the rotary cycle may be determined from the pattern.

X- and Y-Axis Amplifiers

The X and Y amplifiers are identical. Their purpose is to amplify the two 90° out-of-phase voltages furnished by the external two-phase generator. The generator voltages pass through "avc" networks composed of *R4, C2,* and *R22, C8,* before they reach the grids of the X- and Y-amplifier tubes.

These networks are essentially low-pass *R-C* filters with resistance *R* much greater than the capacitive reactance in the frequency range from 5 cps and up (corresponding to generator speeds above 30 rpm). They reduce harmonics in the generator outputs to a greater degree than the fundamentals, producing relatively pure sine-wave voltages at the inputs of amplifiers *V1* and *V4.* The low-pass networks also serve to maintain constant input voltages to the amplifiers over this same frequency range, and to maintain a 90° phase shift between generator voltages and amplifier input voltages.

Z-Axis Circuits

Modulating or timing signals may be connected to the grid of the cathode-ray tube through capacitor *C1.* The high-input resistance provided by *R1* results in negligible loading upon the signal source. A peak-to-peak signal of 45 volts is sufficient to drive the cathode-ray tube between the limits of cutoff and zero bias, so that satisfactory beam blanking can be expected from signals having considerably small amplitude.

Radial Deflection Circuits

Radial deflection of the fluorescent spot on the screen is provided by modulating the screen grids of the deflection amplifiers, *V2, V3* and *V5, V6.* The input to the radial amplifier may be made at terminals *J7* and *J8,* or connector *J10* to a cathode-follower stage (*V7,* 1st section). The output from the cathode follower is amplified by *V7* (2nd section) and fed to both grids of dual-triode *V8.* Outputs from the *V8* plates are fed, in turn, to the screen grids of the X- and Y-axis deflection amplifiers, modulating their outputs accordingly. The radial amplifier requires less than 0.4 rms volts at the input to produce deflection on the screen equal to the circle diameter.

DU MONT MODEL 279

Frequency Response
Vertical Amplifier 10 cps to 100,000 cps, within 1 db
Horizontal Amplifier 10 cps to 50,000 cps, within 1 db
Sweep Circuit 2 to 30,000 cps

Line Rating 115-230 volts, 50-60 cps

Tube Complement

Type	*Function*
6H6 (V1)	Voltage Calibrator
6J5 (V2)	Vertical Input Cathode Follower
6AG7 (V3, V4)	Vertical-Deflection Output
6SN7 (V5)	Single-Sweep Diode, Phase Splitter
6Q5G (V6)	Gaseous Sweep Oscillator
6SL7 (V7)	Sweep Output Cathode Follower
6SN7 (V8)	Flip-Flop Channel *A*
6J5 (V9)	Cathode Follower X_A
6AG7 (V10, V11)	Vertical Output X_A
6J5 (V12)	Cathode Follower X_B
6AG7 (V13, V14)	Horizontal Output X_B
6SN7 (V15)	Sync Input and Single Sweep Bias
6Q5G (V16)	Gaseous Sweep Oscillator
6SL7 (V17)	Sweep Cathode Follower
6SN7 (V18)	Flip-Flop Channel *B*
6J5 (V19)	Cathode Follower Y_B
6AG7 (V20, V21)	Vertical Output Y_B
5SP-A (V22)	Cathode-Ray Tube
5U4G (V23)	Low-Voltage Rectifier
6B4G (V24)	Voltage Regulator
6SJ7 (V25)	Voltage Control Tube
OD3 (V26, V27)	Gaseous Regulators
6X5GT (V28)	Half-Wave Bias Rectifier
3B24 (V29)	High-Voltage Rectifier
2X2A (V30)	High-Voltage Rectifier

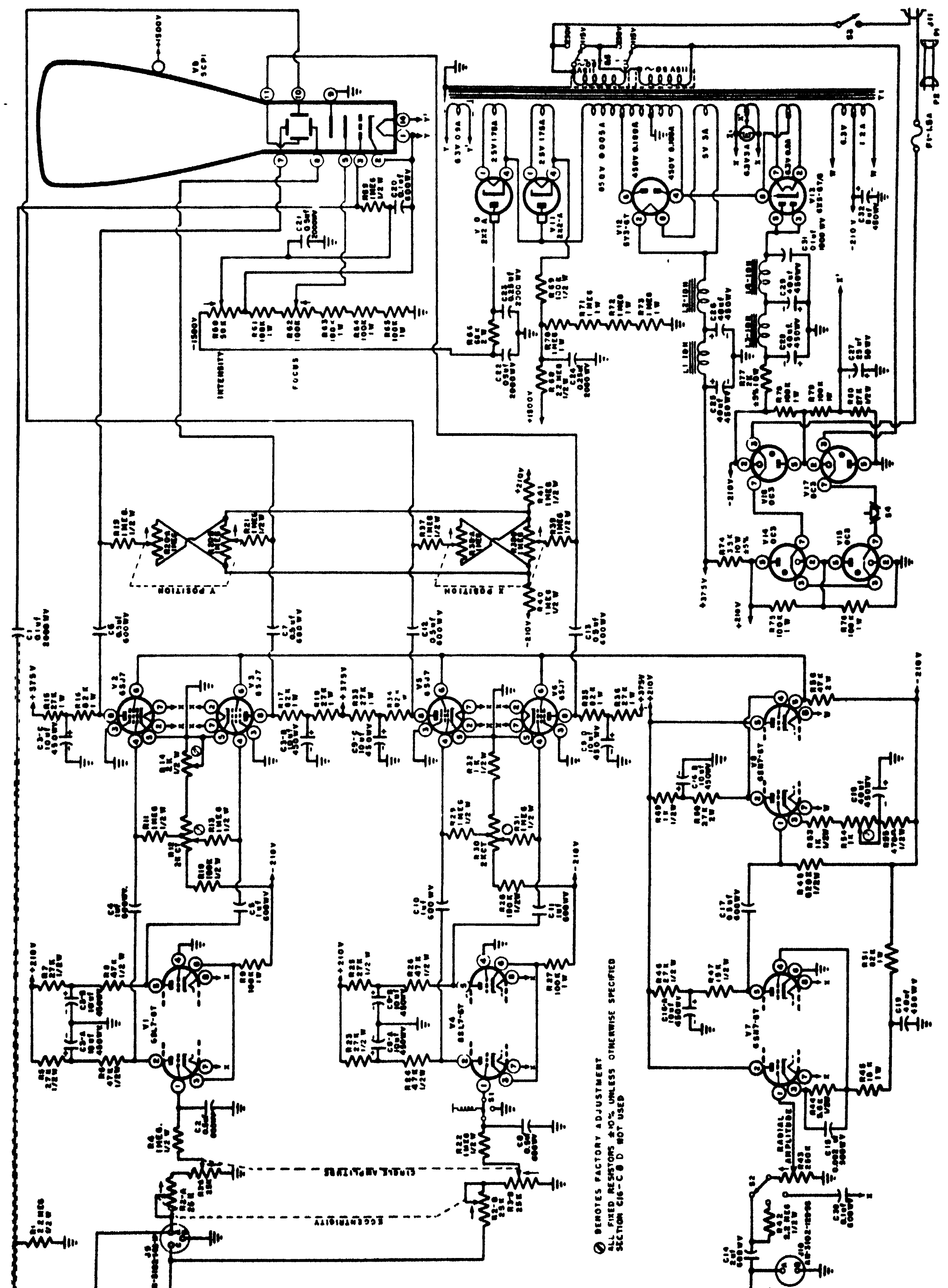
The schematic circuit diagram of Model 279 is shown in Fig. 22-18. This instrument is a specialized type intended for laboratory or scientific work. It is a dual-channel instrument, consisting essentially of two oscilloscopes in one, thus permitting comparison of two waveforms on the same or independent time scales. The instrument utilizes the type 5SP dual beam cathode-ray tube.

Circuits

The two guns and their associated signal deflection circuits are referred to in Fig. 22-18 as Channel *A* and Channel *B,* respectively. The X_A channel handles the *A* beam horizontal deflection and the Y_A the vertical deflection for the *A* beam. Similarly, the *B* beam horizontal and vertical deflection are handled by the corresponding X_B and Y_B channels.

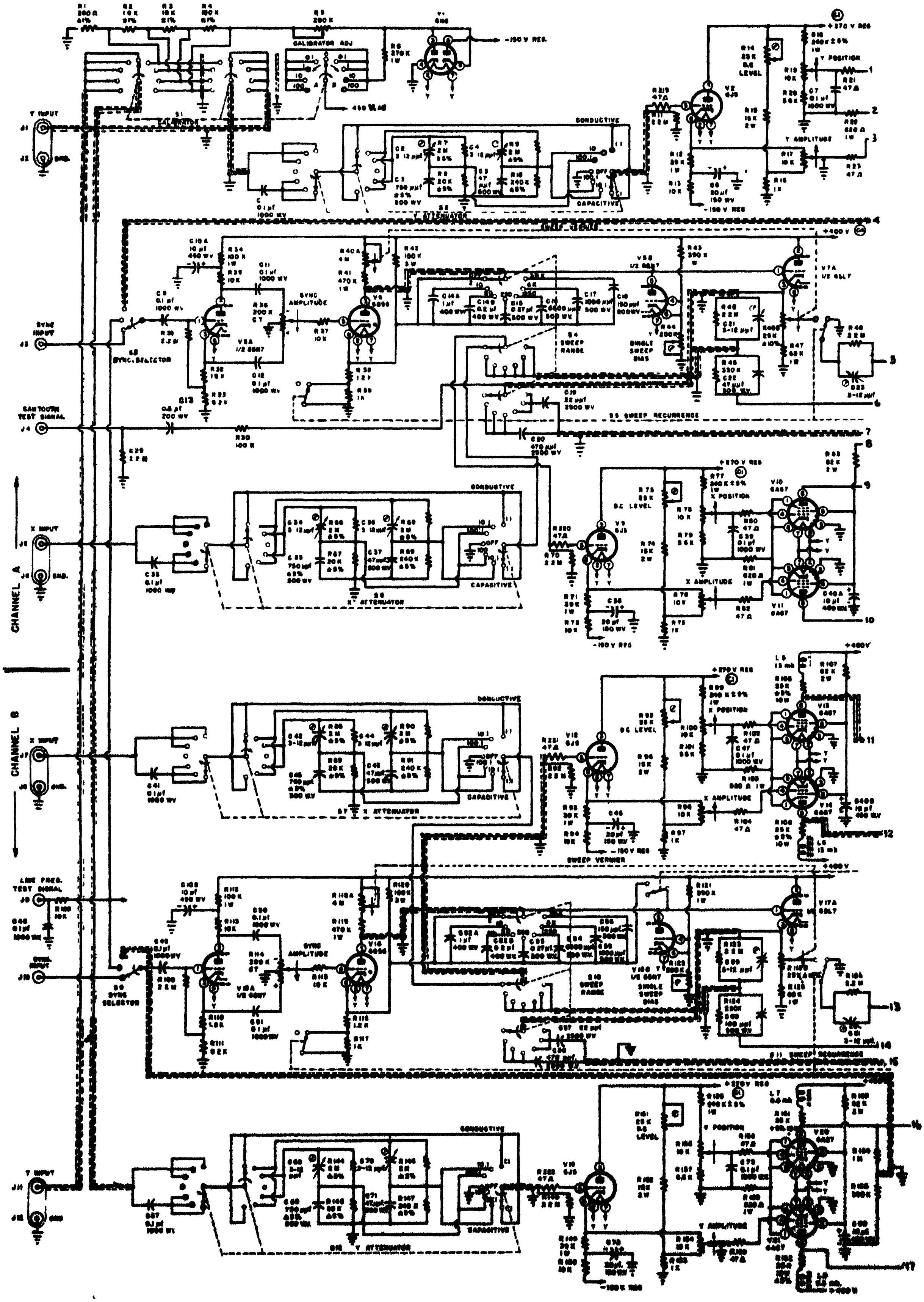
The sync input goes to *V5A* which is a phase splitter. *V6* is a 6Q5G gaseous-triode sweep oscillator. *V5B* is used as a single-sweep diode. The output of the sweep tube is fed to a cathode follower consisting of ½ of a 6SL7, *V7A.*

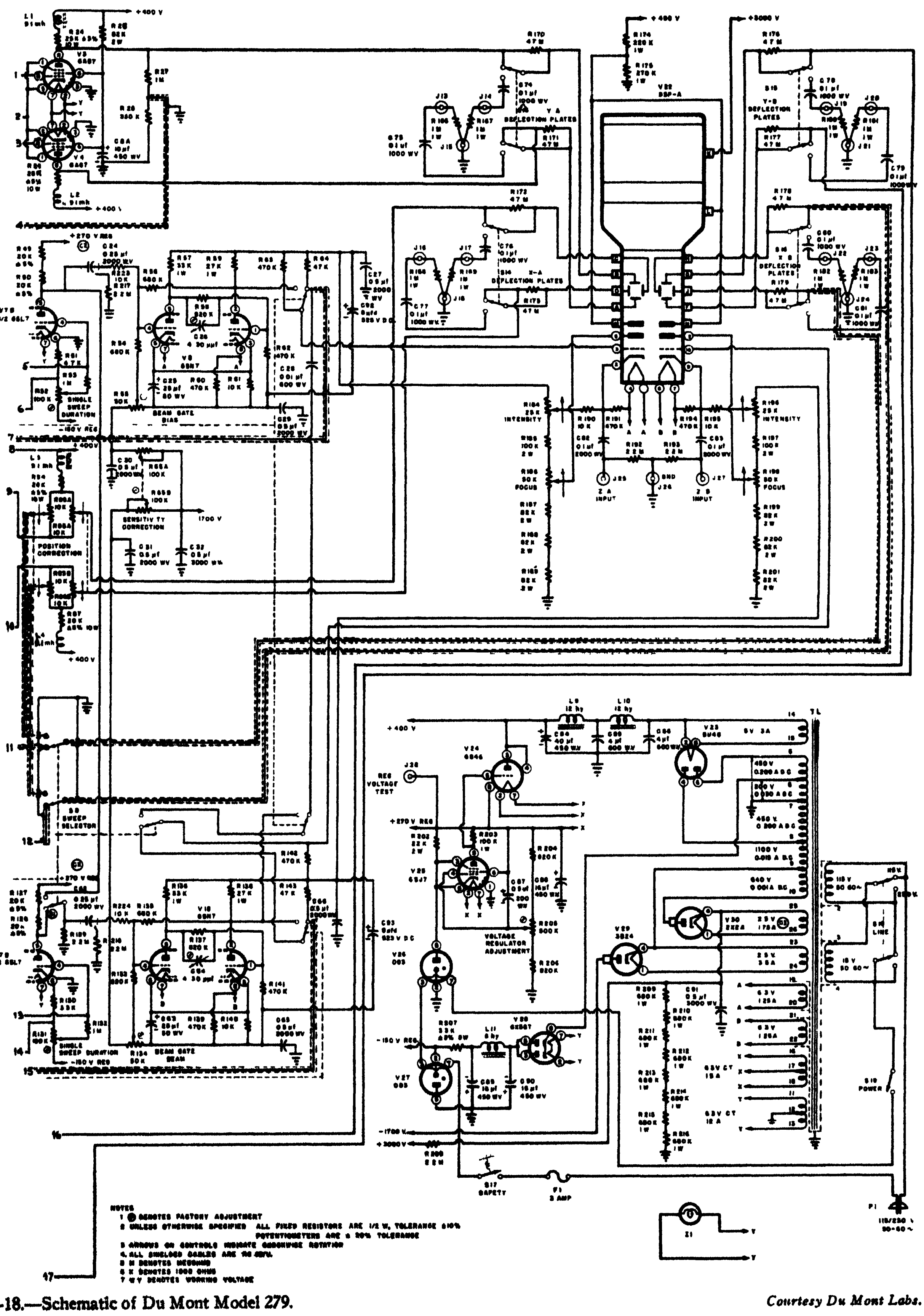
A 6SN7 (*V8* for channel *A, V18* for channel *B*), used in a flip-flop circuit with two stable operating points, is triggered by the signal produced in the plate circuit of *V7B* for sweep *A* (*V17B* for sweep *B*). This flip-flop circuit generates the intensifying gate which illuminates the screen of the cathode-ray tube during the single-sweep period and keeps the tube cut off at all other times. The first half-section is biased off by the BEAM GATE BIAS control and the second half-section is conducting normally. This results in holding the cathode-ray-tube grid potential below the cutoff value. The positive gate signal from the plate of *V7B* turns *V8* on and this, in turn, cuts off *V8B* in whose plate circuit a positive gate is hence generated and fed directly to the grid of the cathode-ray tube, thus turning the beam on. Since the beam-gate circuit is directly coupled throughout, the gate will correspond in duration with the waveform fed into the first grid circuit from the plate of *V7B.*



Courtesy Du Mont Labs.

Fig. 22-17.—Schematic of Du Mont Model 275-A.


Y INPUT
J1
J2
GND.
S1 CALIBRATOR
CALIBRATOR ADJ
V1 6H6
-150 V REG
+400 V REG
CONDUCTIVE
CAPACITIVE
S2 Y ATTENUATOR
V2 6J6
DC LEVEL
Y POSITION
Y AMPLITUDE
+270 V REG
SYNC INPUT
J3
S3 SYNC SELECTOR
V3A 1/2 6SN7
SYNC AMPLITUDE
V4 5656
V5B 1/2 6SN7
V7A 1/2 6SL7
S4 SWEEP RANGE
SINGLE SWEEP BIAS
S5 SWEEP RECURRENCE
SAWTOOTH TEST SIGNAL
J4
CHANNEL A
X INPUT
J5
J6
S6 X ATTENUATOR
V9 6J6
V10 6AG7
V11 6AG7
X POSITION
X AMPLITUDE
CHANNEL B
J7
J8
S7 X ATTENUATOR
V12 6J6
V13 6AG7
V14 6AG7
SWEEP VERNIER
LINE FREQ. TEST SIGNAL
J9
J10
S8 SYNC SELECTOR
V15A 1/2 6SN7
V16 5656
V16B 1/2 6SN7
V17A 1/2 6SL7
S10 SWEEP RANGE
S11 SWEEP RECURRENCE
J11
J12
S12 Y ATTENUATOR
V19 6J6
V20 6AG7
V21 6AG7
+400V
1
2
3
4
5
6
7
8
9
10
11
12
13
14
15
16
17



NOTES
1 ● DENOTES FACTORY ADJUSTMENT
2 UNLESS OTHERWISE SPECIFIED ALL FIXED RESISTORS ARE 1/2 W, TOLERANCE ±10%
POTENTIOMETERS ARE ± 20% TOLERANCE
3 ARROWS ON CONTROLS INDICATE CLOCKWISE ROTATION
4. ALL SHIELDED CABLES ARE [illegible]
5 M DENOTES MEGOHMS
6 K DENOTES 1000 OHMS
7 WV DENOTES WORKING VOLTAGE

Fig. 22-18.—Schematic of Du Mont Model 279.

Courtesy Du Mont Labs.

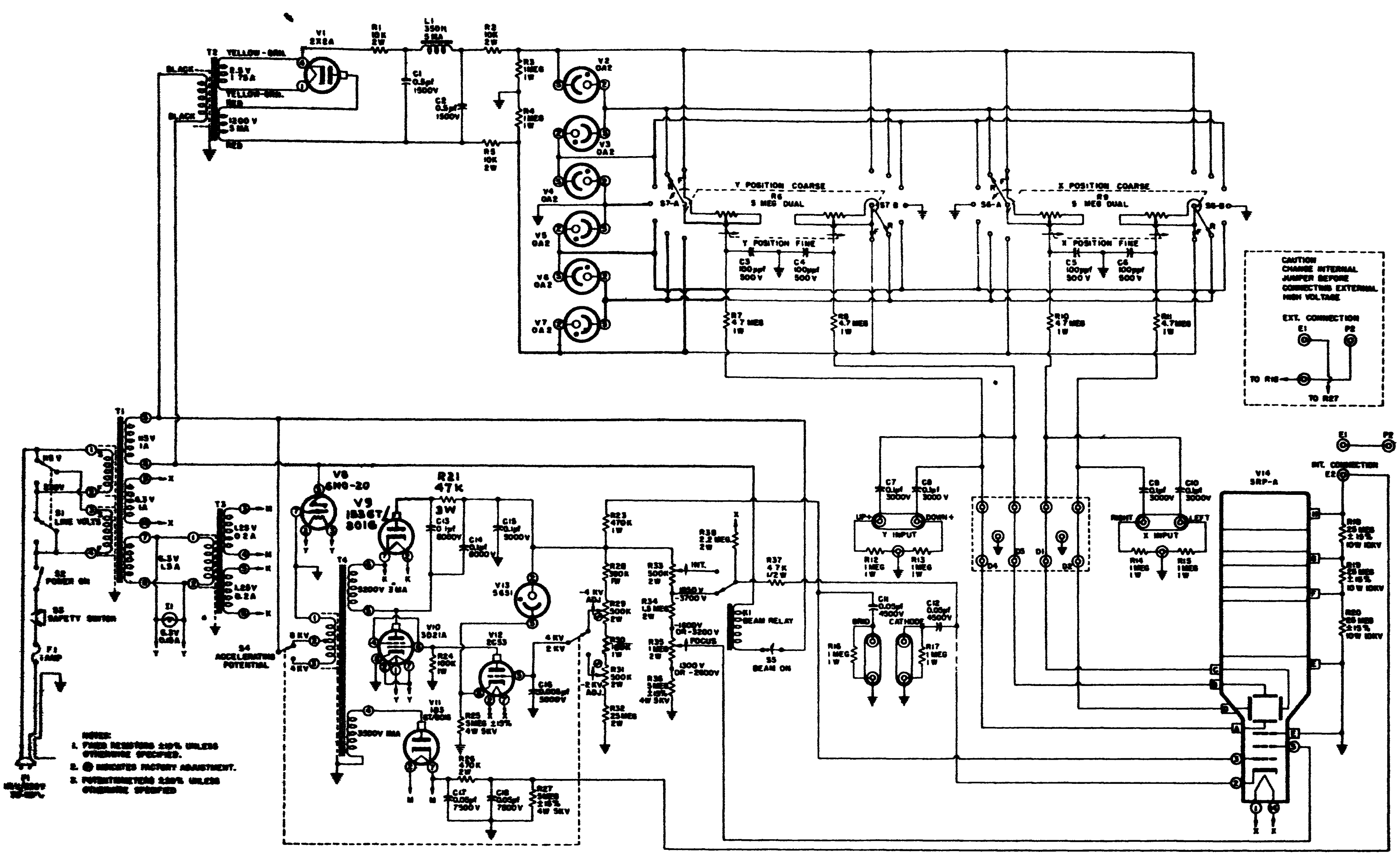


Fig. 22-19.—Schematic of Du Mont Model 281-A.

Courtesy Du Mont Labs.

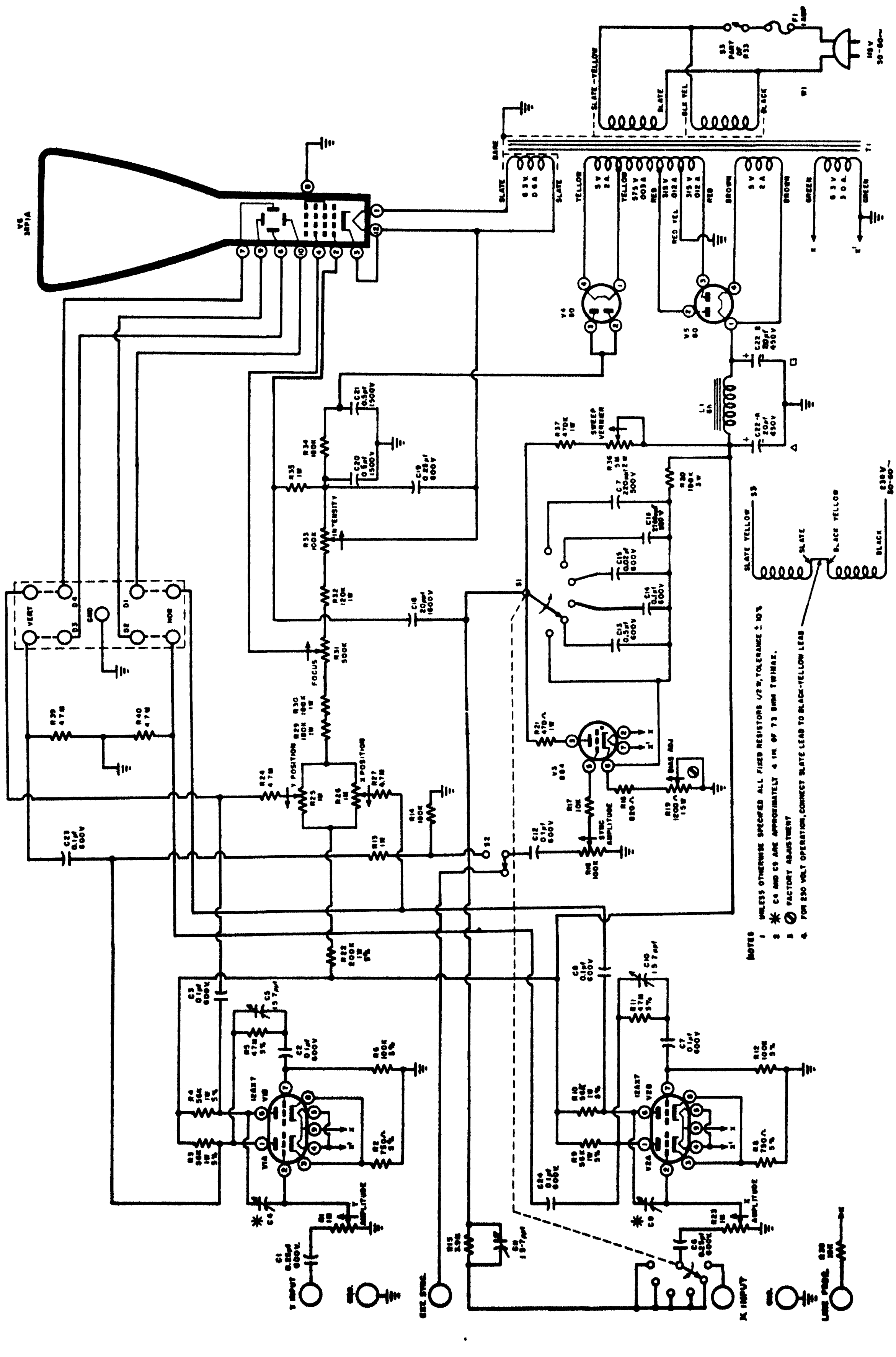
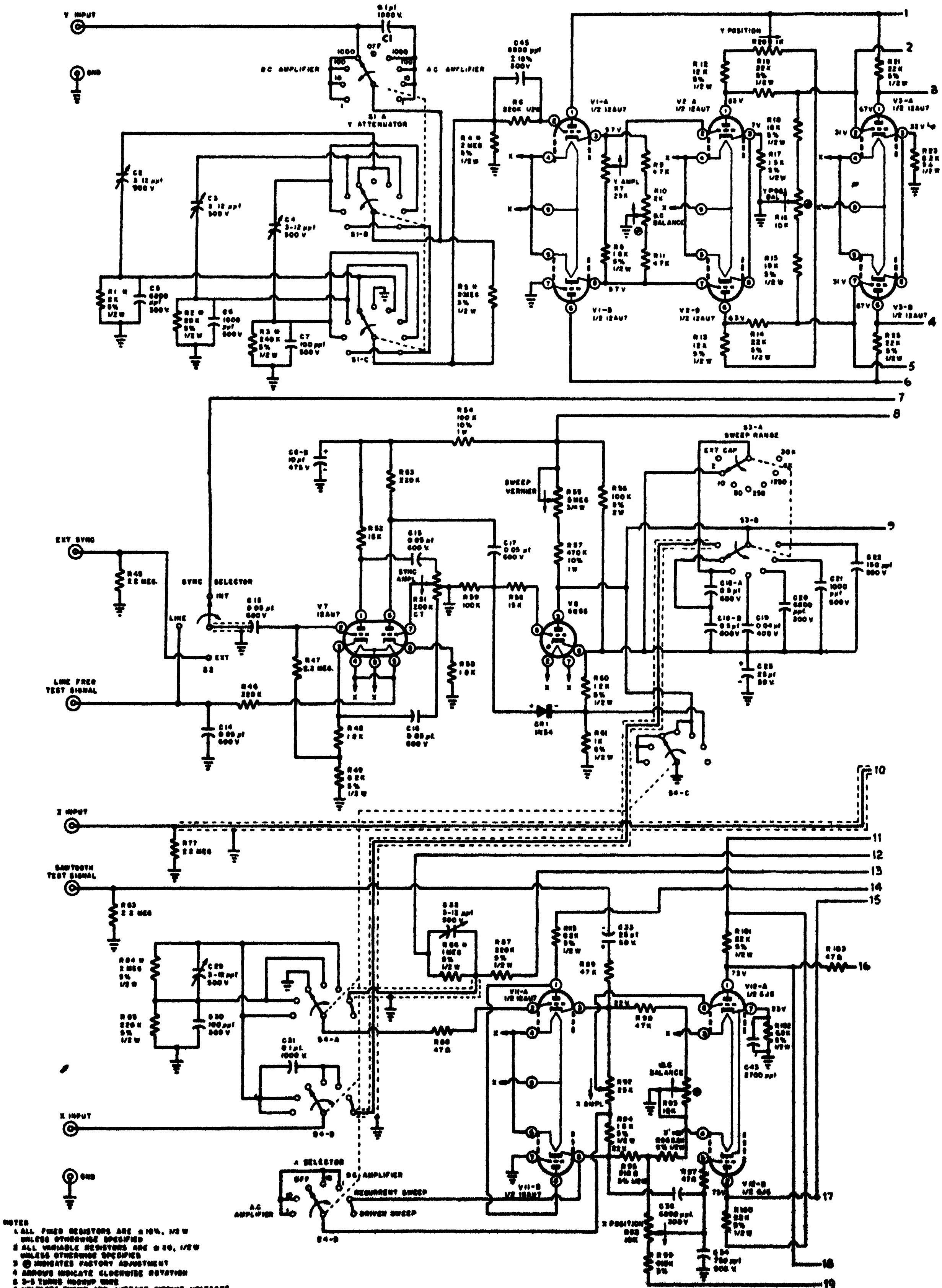

V6 3RP1A
V4 80
V5 80
V3 884
T1
NOTES
1 UNLESS OTHERWISE SPECIFIED ALL FIXED RESISTORS 1/2 W, TOLERANCE ± 10%
2 ✱ C4 AND C9 ARE APPROXIMATELY 4 IN. OF 72 OHM TWINAX.
3 ⊘ FACTORY ADJUSTMENT
4 FOR 230 VOLT OPERATION, CONNECT SLATE LEAD TO BLACK-YELLOW LEAD
Y INPUT
X INPUT
EXT. SYNC
LINE FREQ.
FOCUS
INTENSITY
SWEEP VERNIER
SYNC AMPLITUDE
X POSITION
Y POSITION
X AMPLITUDE
Y AMPLITUDE

Fig. 22-20.—Schematic of Du Mont Model 292.

Courtesy Du Mont Labs.



NOTES

1. ALL FIXED RESISTORS ARE ± 10%, 1/2 W UNLESS OTHERWISE SPECIFIED
2. ALL VARIABLE RESISTORS ARE ± 20, 1/2 W UNLESS OTHERWISE SPECIFIED
3. ● INDICATES FACTORY ADJUSTMENT
4. ARROWS INDICATE CLOCKWISE ROTATION
5. 3-5 TURNS HOOKUP WIRE
6. VOLTAGES SHOWN ARE AVERAGE CIRCUIT VOLTAGES
7. (M) ARE METALLIZED RESISTORS (IRC, TYPE BTS)

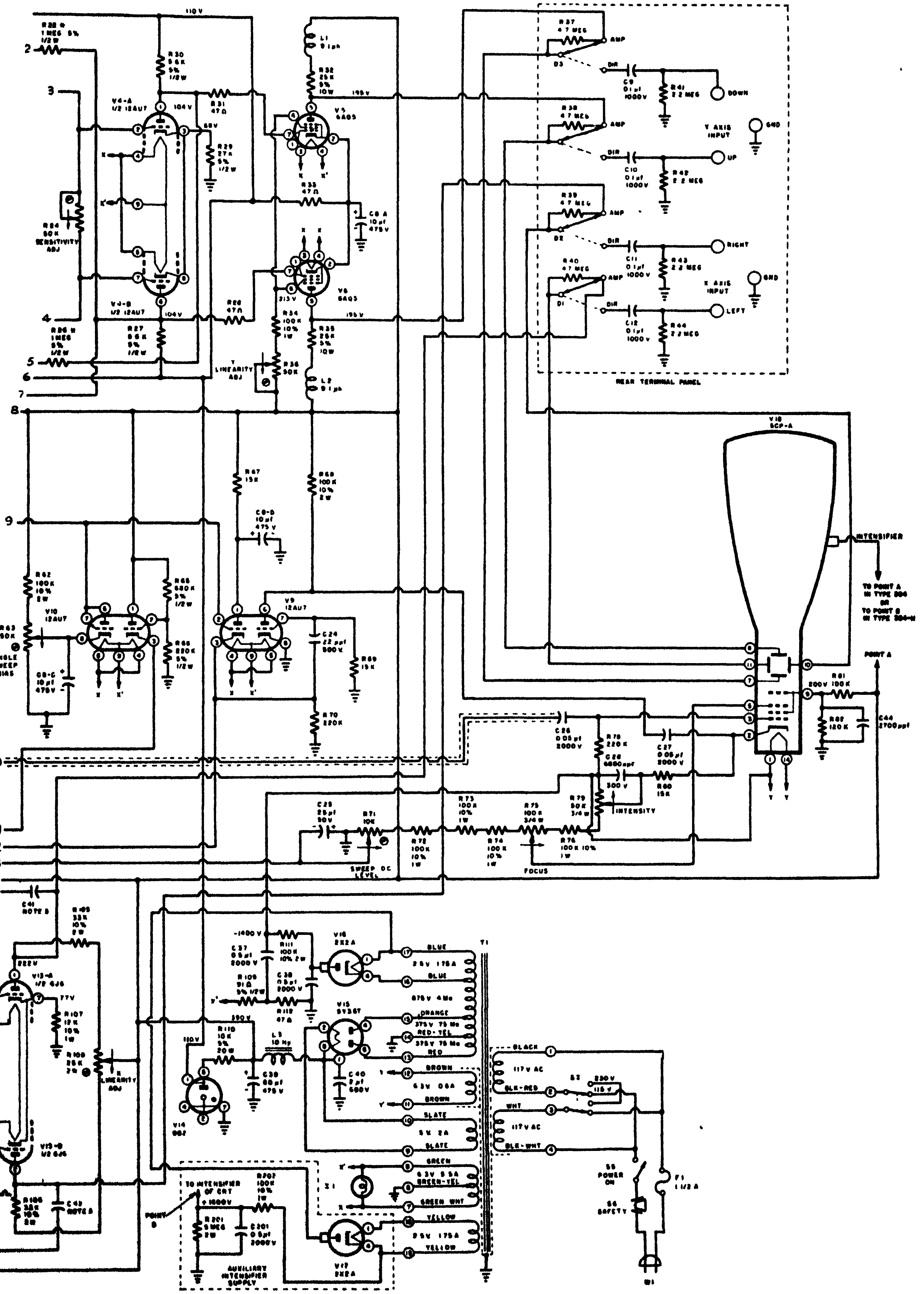


Fig. 22-21.—Schematic of Du Mont Models 304 and 304-H.

Courtesy Du Mont Labs.

DU MONT MODEL 281-A

Deflection Factors
Vertical-Deflection Plates 85 volts/inch
Horizontal-Deflection Plates 90 d-c volts/inch
Line Rating 115-230 volts, 50-60 cps

Tube Complement

Type	*Function*
2X2A (V1)	Positioning Rectifier
OA2 (V2-V7)	Positioning Voltage Regulators
6NO-20 (V8)	Time-Delay Relay
1B3GT/8016 (V9)	Negative Half-Wave Rectifier
3D21A (V10)	Series Regulator
1B3GT/8016 (V11)	Positive Half-Wave High-Voltage Rectifier
2C53 (V12)	Control
5651 (V13)	Regulator Tube
5RP-A (V14)	Cathode-Ray Tube

The schematic circuit diagram of Model 281-A is shown in Fig. 22-19. This instrument is a basic high-voltage indicator unit consisting of a 5RP-A cathode-ray tube and associated circuits. The regulated high-voltage accelerating potentials, including provisions for connection of an external high-voltage supply, produce intense patterns suitable for observation and photographic recording and projection of transients or signals of short duration. Voltage-regulator tubes are also used to stabilize the positioning potentials for the deflection plate circuits.

X- and Y-Axis Circuits

There are no amplifiers in either circuit to provide deflection voltages for the cathode-ray tube. Provision is made for capacitive coupling from front panel terminals to deflection plates through capacitors *C7*, *C8*, *C9* and *C10*, or direct coupling from rear of instrument to all deflection plates from terminals *D1*, *D2*, *D3*, and *D4*.

Z-Axis Circuits

Modulating signals may be fed to the grid and cathode, respectively, of the cathode-ray tube through capacitors *C11* or *C12*. Signals which are not of proper polarity for application to the grid may be applied to the cathode. However, the input impedance to the cathode circiut is much lower than to the grid, and may not be satisfactory for use with some high-impedance signal sources. D-c voltages for the grid and cathode are obtained from the negative high-voltage power supply.

DU MONT MODEL 292

Frequency Response
Vertical Amplifier 5 cps to 100 kc
Horizontal Amplifier 5 cps to 100 kc
Sweep Circuit 8 cps to 30 kc
Deflection Factors
Vertical Amplifier 0.4 rms volts/inch (max)
Vertical-Deflection Plates 22 rms volts/inch
Horizontal Amplifier 0.56 rms volts/inch
Horizontal-Deflection Plates 31 rms volts/inch
Line Rating 115-230 volts, 50-60 cps

Tube Complement

Type	*Function*
12AX7 (V1)	Vertical Amplifier
12AX7 (V2)	Horizontal Amplifier
884 (V3)	Time-Base Generator
80 (V4)	High-Voltage Rectifier
80 (V5)	Low-Voltage Rectifier

The schematic circuit diagram of Model 292 is shown in Fig. 22-20. This basic oscilloscope unit is conventional, except that it is provided with special high-frequency compensation, obtained by means of capacitive feedback (*C4* and *C9*) between plate and grid circuits of the deflection amplifiers.

DU MONT MODELS 304 AND 304-H

Frequency Response
Vertical A-C and D-C Amplifier to 100 kc, within 10%
Horizontal Amplifier A-C and D-C Amplifier to 100 kc, within 10%
Sweep Circuit 2 cps to 30 kc

The sweep is expandable to six times the full-screen diameter of the cathode-ray tubes.

Deflection Factors
Vertical Amplifier 0.01 rms volts/inch
Vertical-Deflection Plates 18 rms volts/inch
Horizontal Amplifier 0.05 rms volts/inch
Horizontal-Deflection Plates 21 rms volts/inch
Line Rating 115 volts, 50-60 cps

Tube Complement

Type	*Function*
12AU7 (V1)	Vertical-Amplifier Input
12AU7 (V2, V3, V4)	Vertical Amplifiers
6AQ5 (V5, V6)	Vertical-Amplifier Output
12AU7 (V7)	Sync Selector
6Q5G (V8)	Timing-Axis Oscillator
12AU7 (V9)	Sweep-Cathode-Follower Blanking Amplifier
12AU7 (V10)	Single-Sweep Clipper
12AU7 (V11)	Horizontal-Amplifier Input
6J6 (V12)	Horizontal Amplifier
6J6 (V13)	Horizontal-Amplifier Output
OB2 (V14)	Voltage Regulator
5Y3 (V15)	Low-Voltage Rectifier
2X2 (V16)	High-Voltage Negative Rectifier
2X2 (V17)	Rectifier Auxiliary Intensifier Supply (304-H only)
5CP-A (V18)	Cathode-Ray Tube

The schematic circuit diagram of Models 304 and 304-H is shown in Fig. 22-21. This is a general-purpose instrument incorporating highly sensitive d-c amplifiers to permit observation of phenomena over an extended frequency range. The following description covers both types 304 and 304-H, which are identical, except that Model 304-H has a supplementary high-voltage rectifier circuit to provide an additional +1,200 volts to the intensifier electrode of the 5CP-A cathode-ray tube.

The positioning circuits of these units are broad enough to permit examination of any portion of the sweep on the cathode-ray-tube screen without distortion. The d-c positioning system permits the equivalent of four times full-scale expansion of the signal for the Y-axis. For the X-axis deflection, the d-c positioning system permits the equivalent of six times full-scale expansion of the signal. The expanded sweep is capable of a sweep writing rate of one inch per microsecond or faster.

Y-Axis Deflection

The signal may be applied to the vertical-deflection plates either directly or through a high-gain amplifier. For low-level signals, the type 304 is provided with a stable, high-gain d-c amplifier. The input system to the Y-amplifier utilizes an attenuator which retains the high-gain d-c characteristics. The input attenuator is of the decade type, using *R-C* compensated, stepped attenuators and a linear gain control *R1*.

The last stage of the amplifier is made up of two type 6AQ5 pentodes. The screens of these pentodes are operated from the unregulated supply, so that the sensitivity of the system rises with increases in line voltage, due to increased amplifier gain. This essentially compensates for the reduction in cathode-ray tube sensitivity, which is experienced at higher accelerating potentials resulting from higher line voltages. As a result of these two compensating factors, the sensitivity of the system (through the amplifier) is almost independent of line-voltage variations.

X-Axis Deflection

For deflection along the X-axis, either the internal sweep signal or an external signal may be used. The X-axis positioning system is somewhat different from that employed on the Y-axis. Here the positioning voltage is applied to one grid of the first amplifier stage, while the input signal is applied to the grid of the other half of this stage. The positioning potentiometer is returned to a point near the cathode end of the cathode load of the cathode follower. This tends to increase amplifier stability, since variations in line voltage now produce about the same variations in voltage at the two grids of the first stage of amplification *V12*.

ELECTRONIC INSTRUMENT CO. (EICO) MODEL 400

FREQUENCY RESPONSE

Vertical Amplifier 30 cps to 50 kc
Horizontal Amplifier 30 cps to 50 kc
Sweep Circuit 30 cps to 30 kc

DEFLECTION FACTORS

Vertical Amplifier 0.35 rms volts/inch
Horizontal Amplifier 0.35 rms volts/inch

LINE RATING 115 volts, 60 cps

TUBE COMPLEMENT

Type	*Function*
6SJ7 (VI)	Vertical-Amplifier Input
884 (V2)	Time-Base Generator
6SJ7 (V3)	Horizontal-Amplifier Input
5Y3 (V4)	Low-Voltage Rectifier
5Y3 (V5)	High-Voltage Rectifier
5BP1 (V6)	Cathode-Ray Tube
6SN7 (V7)	Vertical-Deflection Amplifier
6SN7 (V8)	Horizontal-Deflection Amplifier

The schematic circuit diagram of Model 400 is shown in Fig. 22-22.

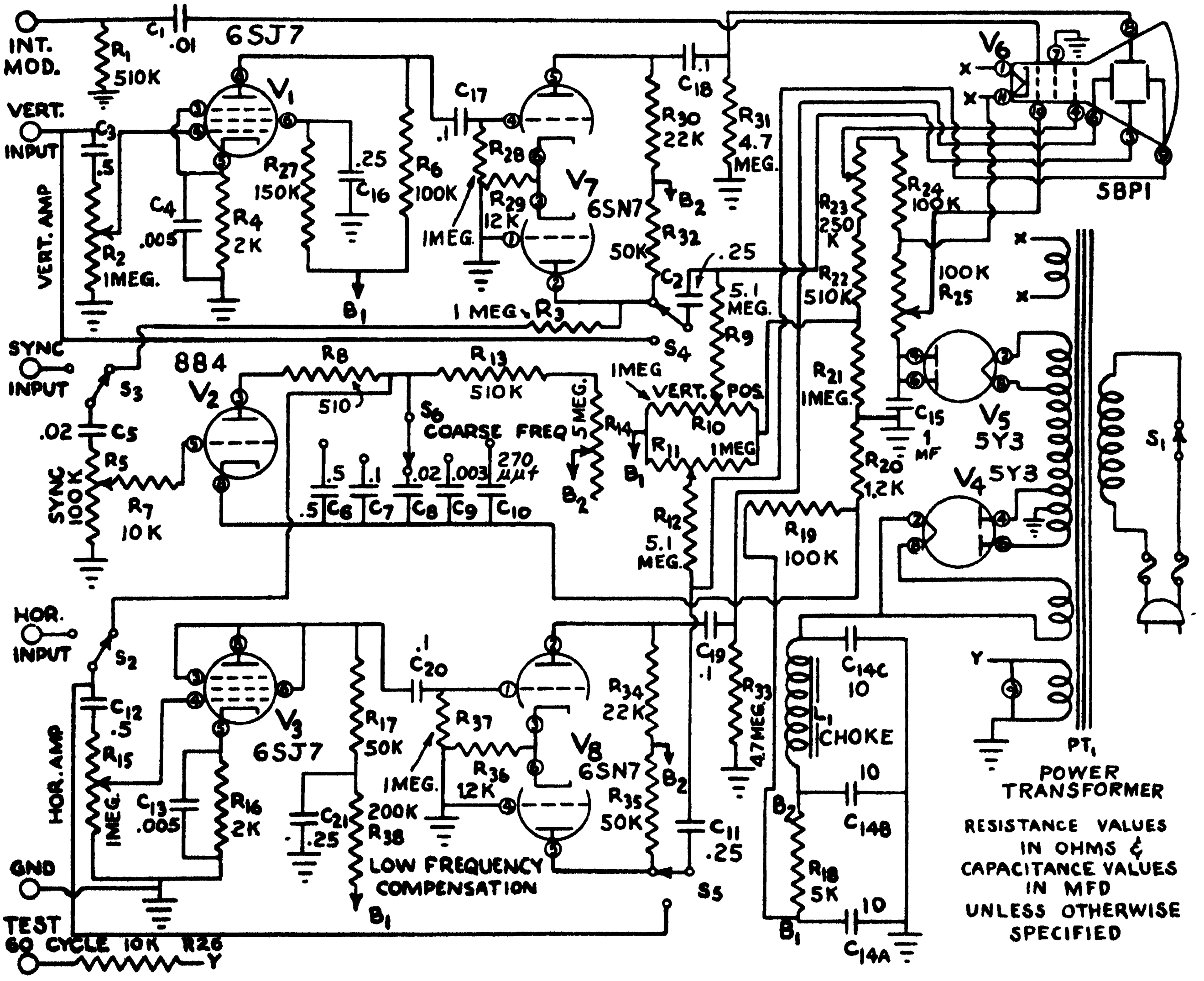


Fig. 22-22.—Schematic of Eico 400.

Courtesy Electronic Inst. Corp.

ELECTRONIC INSTRUMENT CO. (EICO) MODEL 425

FREQUENCY RESPONSE
Vertical Amplifier 5 cps to 400 kc
Horizontal Amplifier 5 cps to 400 kc
Time-Base Generator 15 cps to 75 kc

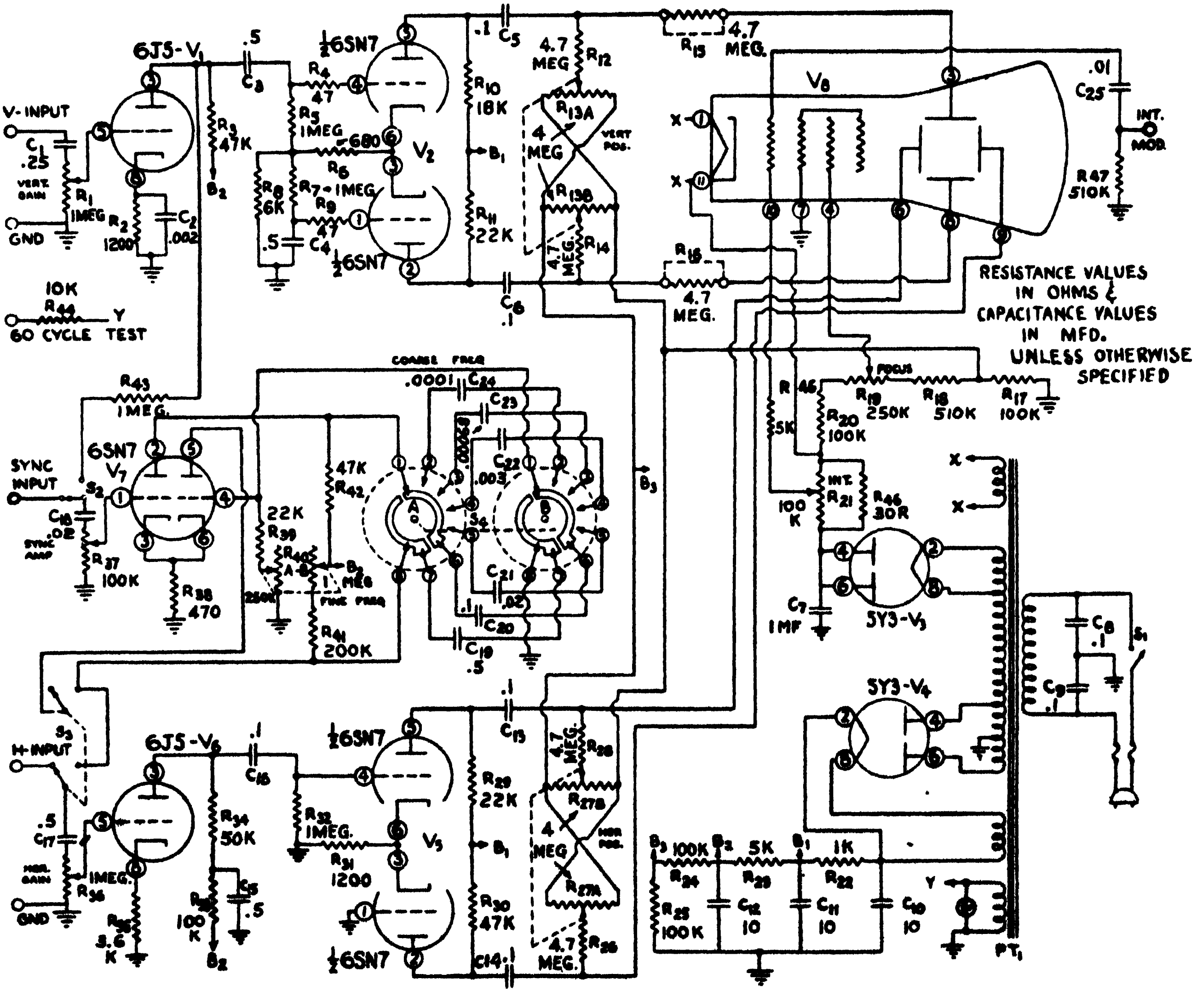
DEFLECTION FACTORS
Vertical Amplifier 0.05 rms volts/inch
Horizontal Amplifier 0.05 rms volts/inch

LINE RATING 100-130 volts, 50-60 cps

TUBE COMPLEMENT

Type	*Function*
6J5 (V1)	Vertical-Amplifier Input
6SN7 (V2)	Vertical-Deflection Amplifier
5Y3 (V3)	High-Voltage Rectifier
5Y3 (V4)	Low-Voltage Rectifier
6SN7 (V5)	Horizontal-Deflection Amplifier
6J5 (V6)	Horizontal-Amplifier Input
6SN7 (V7)	Time-Base Generator
5BP1 (V8)	Cathode-Ray Tube

The schematic circuit diagram of Model 425 is shown in Fig. 22-23.



RESISTANCE VALUES IN OHMS & CAPACITANCE VALUES IN MFD. UNLESS OTHERWISE SPECIFIED

Fig. 22-23.—Schematic of Eico 425.

Courtesy Electronic Inst. Corp.

ELECTRONIC TUBE CORP. MODEL H-21

FREQUENCY RESPONSE
Vertical and Horizontal Amplifiers
Capacitive Input 10 cps to 100 kc, ± 1 db
Conductive Input d.c. to 100 kc, 1 db down
Sweep Circuit 2 cps to 50 kc
Triggered Sweep ½ second to 20 microseconds

DEFLECTION FACTORS
Horizontal Amplifiers 0.75 d-c volts/inch
Direct to Horizontal Plates 69 d-c volts/inch
Vertical Amplifiers 0.1 d-c volts/inch
Direct to Vertical Plates 80 d-c volts/inch

LINE RATING 115 volts, 50-60 cps

The schematic circuit diagram of Model H-21 is shown in Fig. 22-24. The instrument is particularly suited for simultaneous comparison of two electrical waveshapes in amplitude, phase, or frequency, since it contains a dual-gun cathode-ray tube. Four amplifiers, two horizontal and two vertical, are provided for the two channels. Each amplifier is independent, having separate gain adjustments. Either capacitive or conductive input coupling may be used with any of the amplifiers; differential input can also be used with the Y-amplifiers.

The intensity of either or both cathode-ray-tube beams may be modulated by time marker pulses if desired. A blanking switch for suppressing the retrace of the cathode-ray beam is provided. The circuits employed in the H-21 are of conventional design.

FEILER MODEL TS-7A

FREQUENCY RESPONSE
Vertical Amplifier 20 cps to 100 kc, ± 20%
Horizontal Amplifier 20 cps to 100 kc, ± 20%
Sweep Circuit 10 cps to 35 kc

DEFLECTION FACTORS
Vertical Amplifier 0.5 rms volts/inch
Vertical-Deflection Plates 18 rms volts/inch, ± 20%
Horizontal Amplifier 0.5 rms volts/inch
Horizontal-Deflection Plates 18 rms volts/inch, ± 20%

LINE RATING 105-125 volts, 50-60 cps

The Feiler TS-7A "Stethoscope" consists of a special signal-tracing probe and a standard oscilloscope. The circuits are conventional and are shown in the schematic diagram, Fig. 22-25. The probe uses a *6C4* grid-leak detector which has only slight loading effect on the circuit under test due to the use of a very small coupling capacitance in the probe tip circuit.

GENERAL ELECTRIC MODEL CRO-3A

FREQUENCY RESPONSE
Vertical Amplifier 20 cps to 100 kc, ±10%
Horizontal Amplifier 20 cps to 100 kc, ±10%
Sweep Circuit 20 cps to 30 kc

DEFLECTION FACTORS
Vertical Amplifier 0.3 rms volts/inch
Vertical-Deflection Plates 22 d-c volts/inch
Horizontal Amplifier 0.3 rms volts/inch
Horizontal-Deflection Plates 20 d-c volts/inch

LINE RATING 105-125 volts, 50-60 cps

TUBE COMPLEMENT

Type	*Function*
3AP1A (V1)	Cathode-Ray Tube
5Y3GT (V2)	High-Voltage Rectifier
5Y3 (V3)	Low-Voltage Rectifier
6AC7 (V4)	Vertical Amplifier
6AC7 (V5)	Horizontal Amplifier
884 (V6)	Sweep Oscillator

The schematic circuit diagram of Model CRO-3A is shown in Fig. 22-26. This is a portable, general-purpose oscilloscope.

GENERAL ELECTRIC MODEL CRO-5A

FREQUENCY RESPONSE
Vertical Amplifier 20 cps to 350 kc, 10% down
Horizontal Amplifier 20 cps to 200 kc, 10% down
Sweep Circuit 20 cps to 20 kc

DEFLECTION FACTORS
Vertical Amplifier 0.02 rms volts/inch
Vertical-Deflection Plates 33 d-c volts/inch
Horizontal Amplifier 0.2 rms volts/inch
Horizontal-Deflection Plates 31 d-c volts/inch

LINE RATING 105-125 volts, 60 cps

The schematic circuit diagram of Model CRO-5A is shown in Fig. 22-27. This instrument is intended for general-purpose test and development work. The vertical-input stage is a cathode follower which functions as an impedance transformer and cuts down the input-capacitive reactance. *V2, V3* and *V4* use adjustable plate-load inductances for high-frequency compensation. The grid excitation circuit for *V4* is somewhat unusual. A portion of the signal plate voltage at *V3* is picked off through *R15, R16*, and *C12*, and is applied to the *V4* grid approximately at 180° out of phase with the signal on the *V3* grid due to the phase shift in *V3*. Tubes *V3* and *V4*, therefore, act as push-pull output tubes. A similar circuit is used in the horizontal amplifier.

GENERAL ELECTRIC MODEL ST-2A

FREQUENCY RESPONSE
Vertical Amplifier (A.C.) (with probe) 20 cps to 500 kc —20%
Vertical Amplifier (D C.) 0 cps to 500 kc, —20%
Horizontal Amplifier (D.C.) 0 cps to 100 kc, —20%
Sweep Circuit 10 cps to 100 kc

DEFLECTION FACTORS
Vertical Amplifier 0.035 rms volts/inch
Vertical Probe 0.45 rms volts/inch
Horizontal Amplifier 0.425 rms volts/inch

LINE RATING 105-125 volts, 50-60 cps

The schematic circuit diagram for Model ST-2A is shown in Fig. 22-28. The ST-2A is intended for general-purpose use in laboratories, but in addition has several features which make it applicable particularly to f-m and television servicing.

GENERAL ELECTRIC MODEL YNA-4

FREQUENCY RESPONSE
Vertical Amplifier 0 to 30 kc, within + 0% —15%
Horizontal Amplifier 0 to 30 kc
Sweep Circuit 4 cps to 20 kc

DEFLECTION FACTORS
Vertical Amplifier 0.18 rms volts/inch, max
Horizontal Amplifier 0.212 rms volts/inch, max

LINE RATING 105-125 volts, 50-60 cps

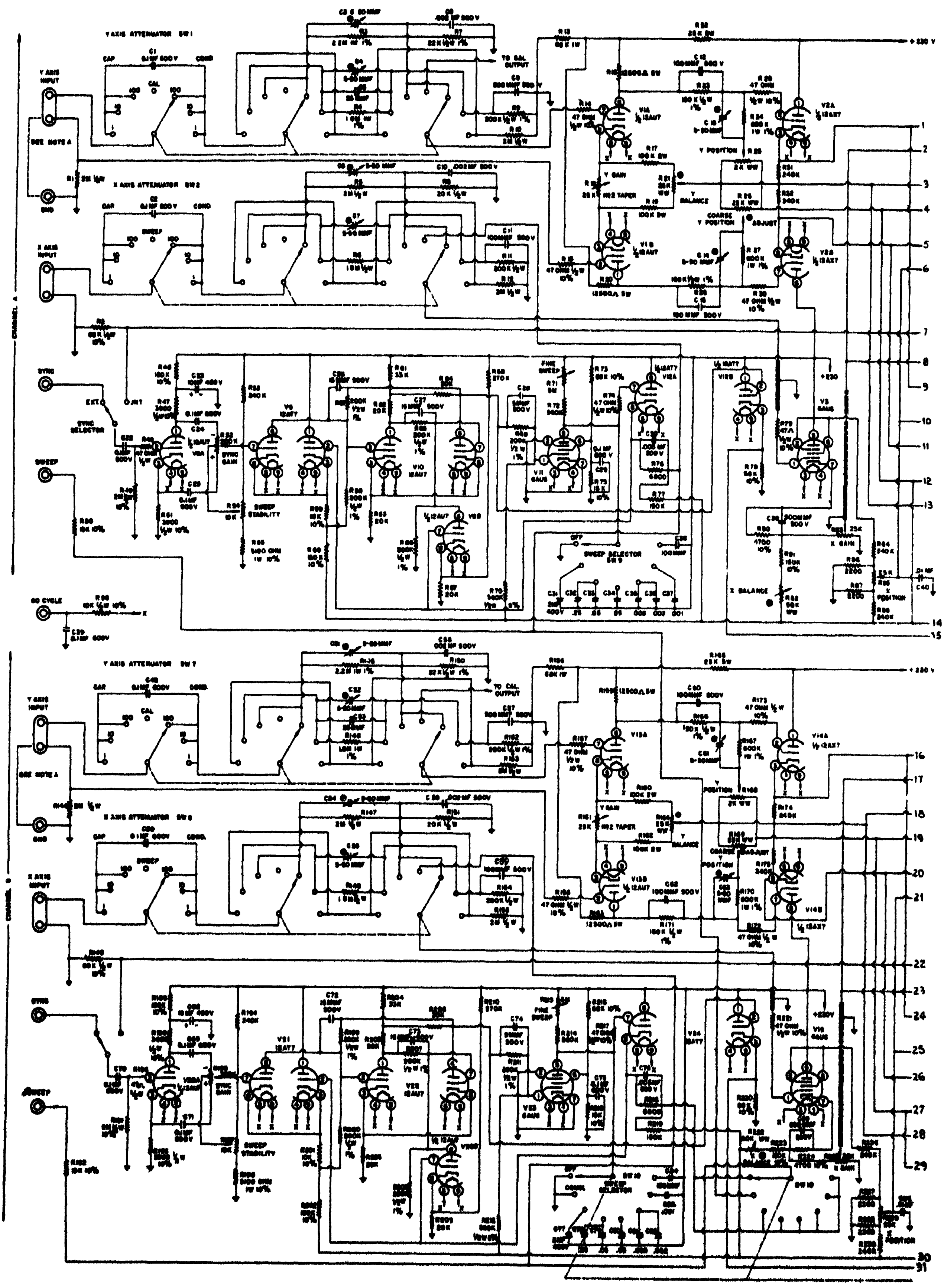
The schematic circuit diagram for Model YNA-4 is shown in Fig. 22-29. This oscilloscope is a portable unit designed specifically for service applications on industrial electronic equipment such as welding controls, motor control circuits, servo mechanisms, photoelectric circuits, etc.

Deflection Amplifiers

The deflection stages are somewhat unusual. *V1* is the vertical-amplifier input tube which, with *V2*, forms the push-pull vertical amplifier. *V2* is a grounded-grid amplifier. If a positive voltage is applied to the grid of *V1*, its cathode, as well as the cathode of *V2* to which it is tied, will go more positive. This effectively *decreases* the grid potential on *V2*, since its grid is grounded and cannot change. Thus a push-pull effect is obtained from plates of *V1* and *V2* which is applied to the deflection plates of the cathode-ray tube.

Centering Controls

R7A and *R7B* form a ganged vertical-centering control, to vary the screen potentials and, therefore, the plate potentials of *V1* and *V2*. The plates are directly coupled to the 3KP1 deflection plates. When the screen of *V1* is made more or less positive with respect to ground by adjustment of *R7A*, the *V1* plate becomes more or less positive with respect to ground, and hence the spot on the screen is moved. A similar action occurs for *V2* and *R7B*. If the *V2* plate has a higher positive potential to ground than the *V1* plate, the *V1* plate is negative with respect to *V2* and the spot on the cathode-ray-tube screen moves away from the deflection plate connected to the *V1* plate. Thus, the spot may be moved up or down on the vertical axis. The horizontal positioning is controlled similarly by *R20A* in conjunction with *V3*, and *R20B* in conjunction with *V4*.


Y AXIS ATTENUATOR SW1
Y AXIS INPUT
X AXIS ATTENUATOR SW2
X AXIS INPUT
CHANNEL A
SYNC
SWEEP
SYNC SELECTOR
SWEEP STABILITY
SWEEP SELECTOR SW9
FINE SWEEP
TO CAL OUTPUT
Y GAIN
Y POSITION
Y BALANCE
COARSE Y POSITION
ADJUST
X GAIN
X BALANCE
X POSITION
60 CYCLE
Y AXIS ATTENUATOR SW7
X AXIS ATTENUATOR SW8
CHANNEL B
SEE NOTE A
+230 V
1
2
3
4
5
6
7
8
9
10
11
12
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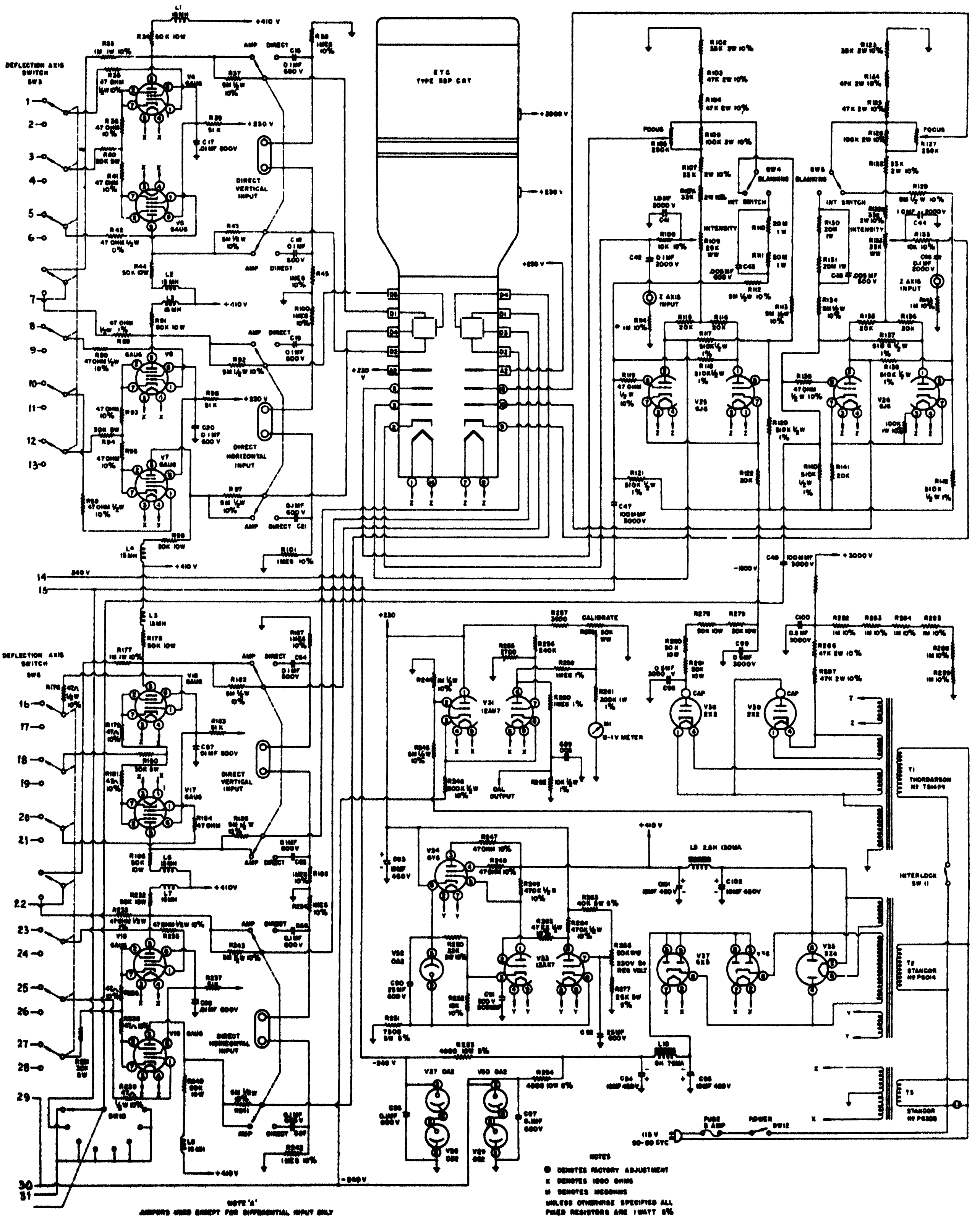


Fig. 22-24.—Schematic of Electronic Tube Model H-21.

Courtesy Electronic Tube Corp.

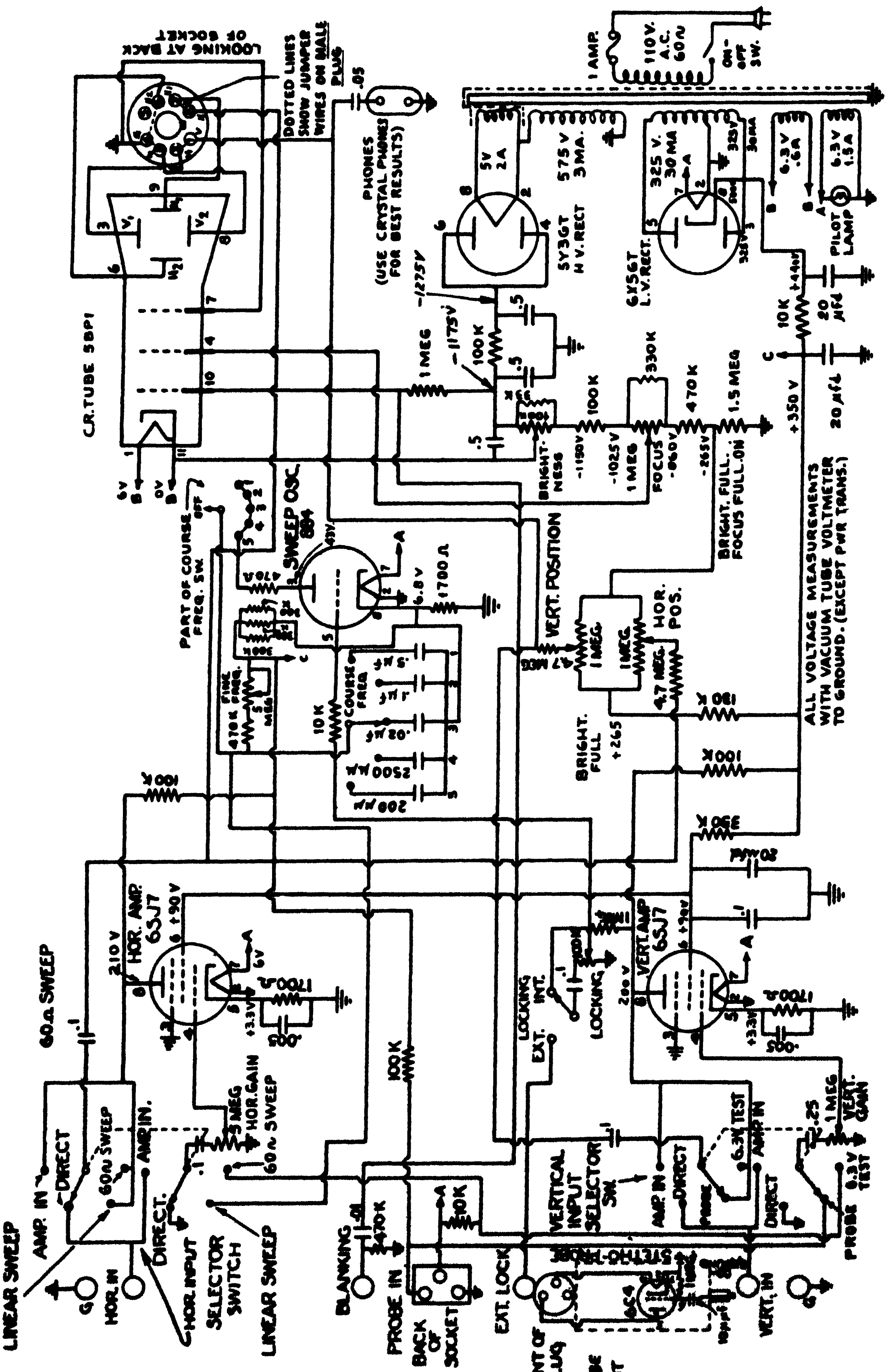
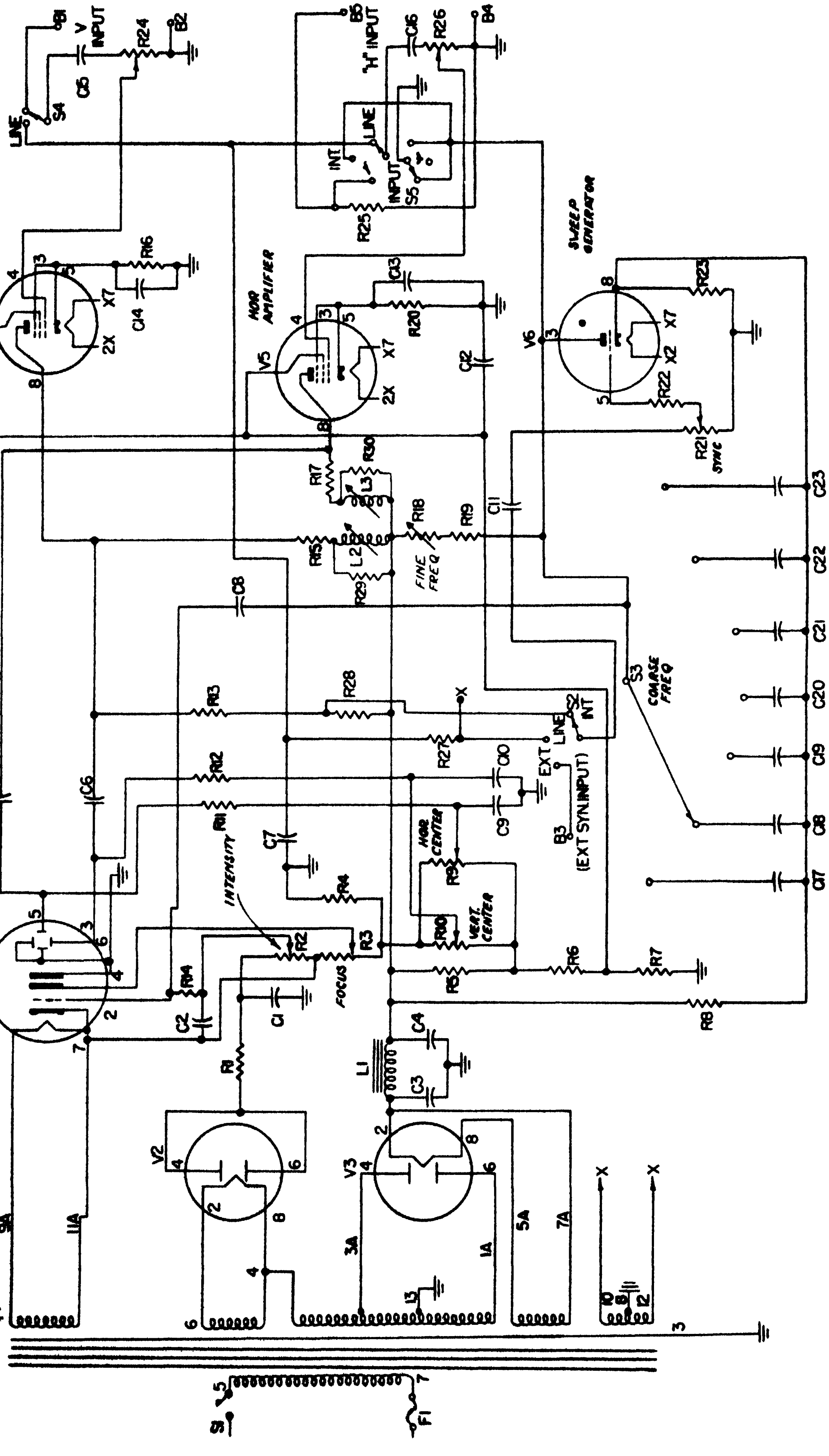


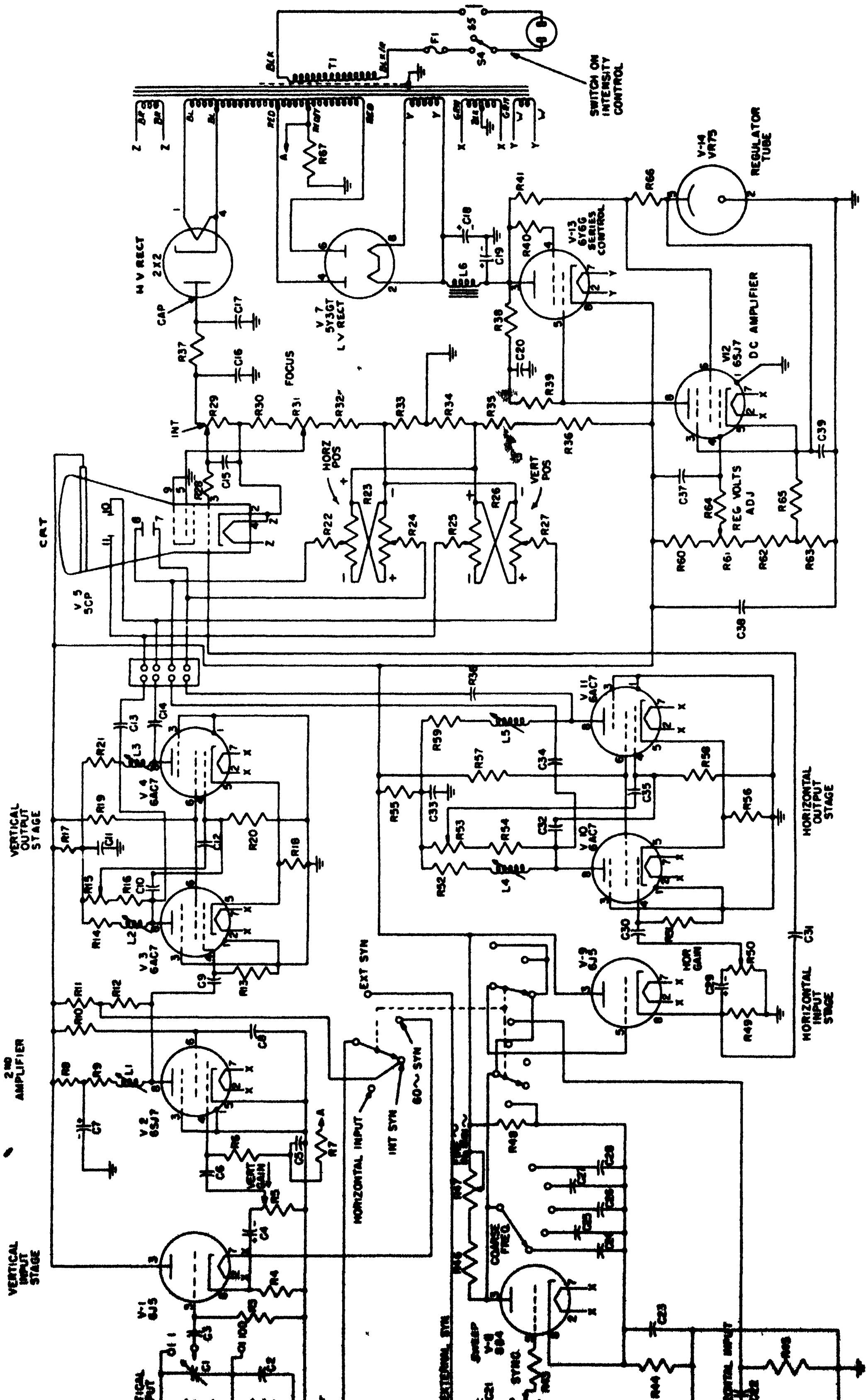
Fig. 22-25.—Schematic of Feiler Model TS-4A.

Courtesy Feiler Eng. Co.



Courtesy General Electric

Fig. 22-26.—Schematic of G.E. Model CRO-3A.



Courtesy General Electric

Fig. 22-27.—Schematic of G.E. Model CRO-5A.

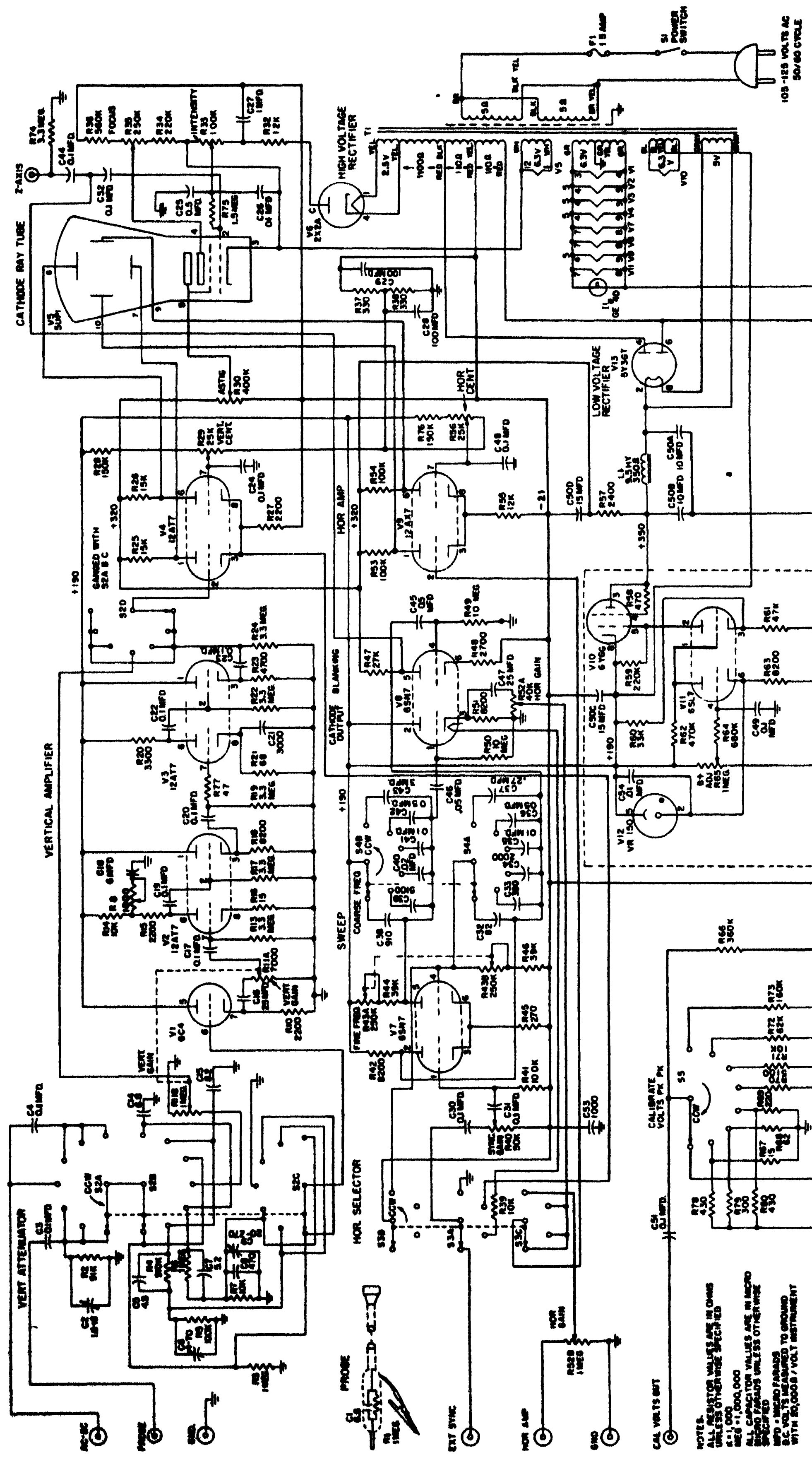
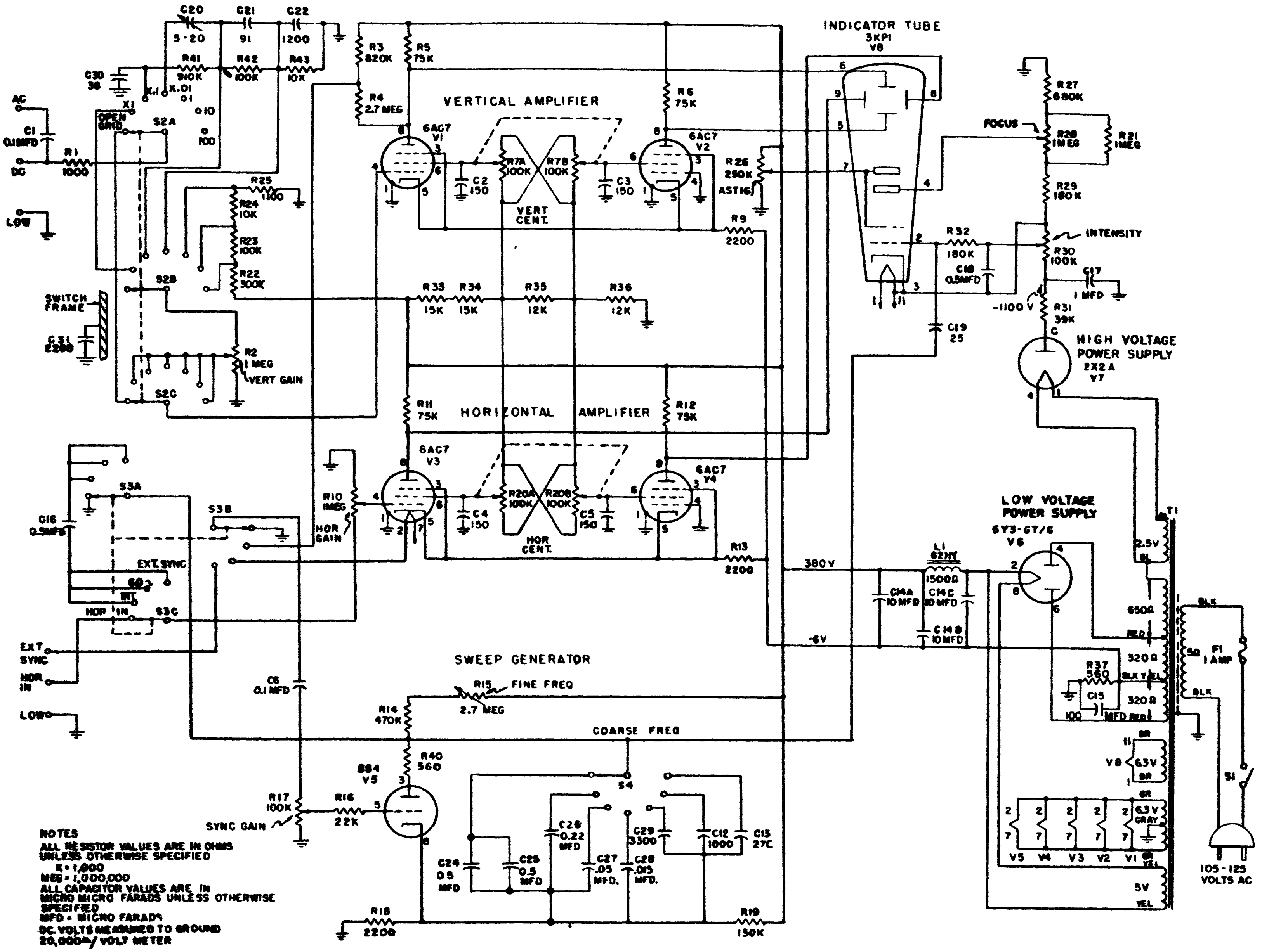


Fig. 22-28.—Schematic of G E. Model ST-2A.

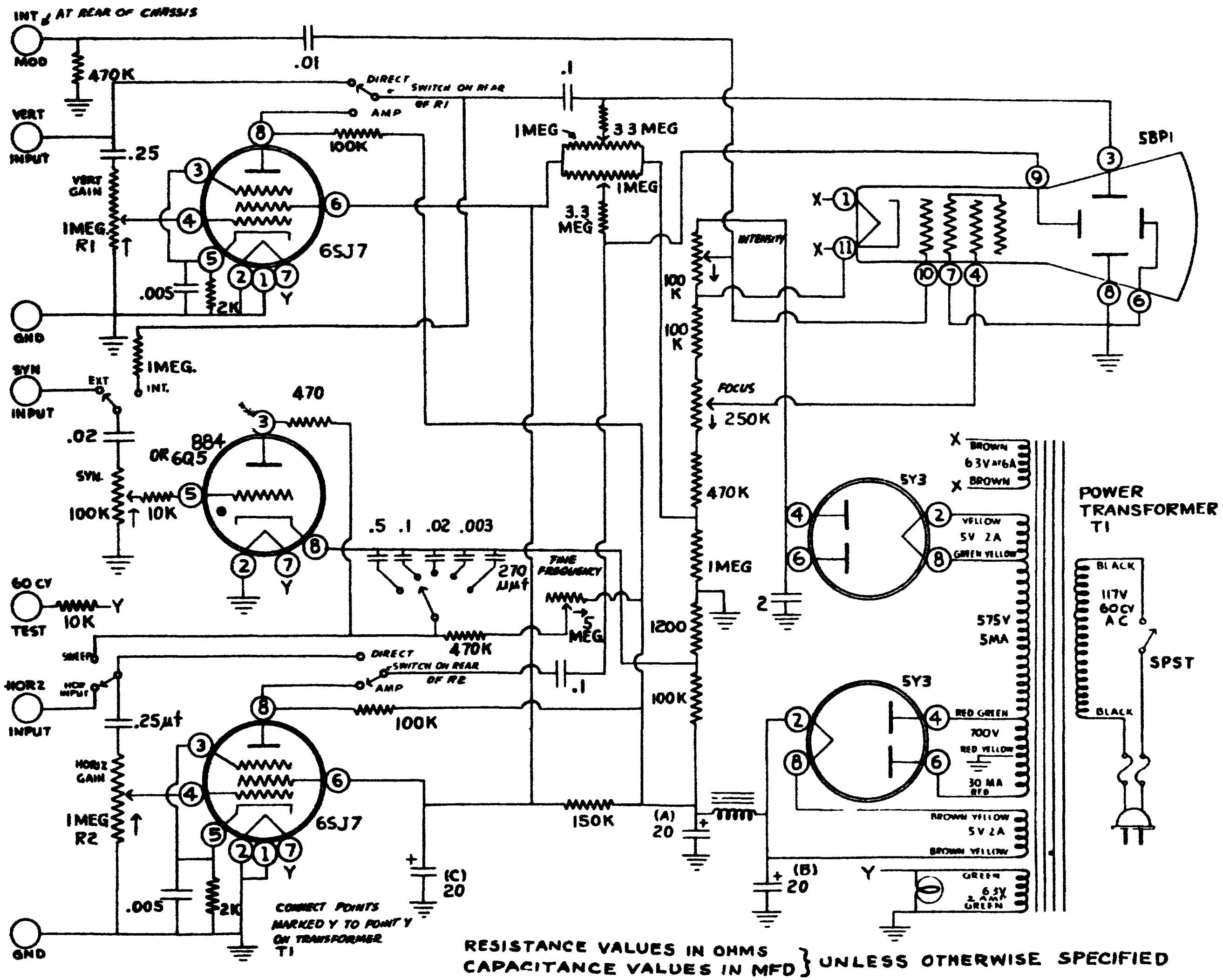
Courtesy General Electric



NOTES
ALL RESISTOR VALUES ARE IN OHMS UNLESS OTHERWISE SPECIFIED
K = 1,000
MEG = 1,000,000
ALL CAPACITOR VALUES ARE IN MICRO MICRO FARADS UNLESS OTHERWISE SPECIFIED
MFD = MICRO FARADS
DC VOLTS MEASURED TO GROUND 20,000Ω/VOLT METER

Fig. 22-29.—Schematic of G.E. Model YNA-4.

Courtesy General Electric



Courtesy Heath Co.

Fig. 22-30.—Schematic of Heath Model O-4.

HEATH CO. MODEL O-4

Line Rating 117 volts, 60 cps

Tube Complement

Type	*Function*
884	Sweep Oscillator
6SJ7	Vertical Amplifier
6SJ7	Horizontal Amplifier
5Y3	High-Voltage Rectifier
5Y3	Low-Voltage Rectifier
5BP1	Cathode-Ray Tube

The schematic circuit diagram of Model 0-4 is shown in Fig. 22-30. This is an oscilloscope construction kit. When properly assembled, it may be used as a general-purpose oscilloscope.

HICKOK MODEL RFO-5

FREQUENCY RESPONSE
  Sweep Circuit (linear) 10 cps to 25 kc
  F-M Sweep Width 10-30 kc, and 100-900 kc
DEFLECTION FACTORS
  Vertical Amplifier (Standard) 0.2 rms volts/inch
  Vertical Amplifier Video 4 rms volts/inch
  Horizontal Amplifier 0.3 rms volts/inch
  Horizontal-Deflection Plates 43.4 volts/inch
  Vertical Plates 56.4 volts/inch
LINE RATING 105-130 volts, 50-60 cps

The schematic circuit diagram for Model RFO-5, including tube complement and tube functions, is shown in Fig. 22-31. This instrument is a combined oscilloscope and sweep generator.

The vertical-input system is quite interesting. The excitation for the 6J7 (*V103*) grid is obtained through *S102, R114,* and *C109* from *H102. R114* is the vertical-gain control. The vertical-input terminal circuit *H102* may also be switched from the *V103* stage to the *V107* video amplifier stage. Demodulator *V106* is switched into the circuit automatically and consists of a simple diode rectifier. With the switch *S102* set to video, the horizontal plates of the cathode-ray tube are fed signal voltage through *C108* from *V106.* The arm of *R114* then connects to *R117* and the *V106* grid circuit.

A rather unusual and novel feature is the incorporation of a jack, *J101,* for signal tracing. An earphone, audio-amplifier input, or a signal tracer can be plugged into *J101.* If a headphone set is used, it should be preferably a high-impedance type, as should any device connected into the *J101* circuit.

The R.F. OSC. MIXER section comprises the remainder of the circuit. The 6K8 section, pins 8-5-6, *V107,* forms a triode oscillator of the Hartley type. The plate connects to one terminal of the 1,000 kc oscillator coil, but may be shifted to the 23-Mc coil *L101* by varying the switch setting in the oscillator coil circuit. Because of the internal direct connection of the triode grid to the 6K8-pentode 1st grid, there is a modulation of the eletron stream in the left-hand portion of the 6K8 as the frequency of the triode oscillator takes place. This electron stream may also be modulated by a signal voltage appearing across *R138,* of 7,500 ohms, derived from the EXT. OSC. INPUT terminal *H107.* The output of the oscillator is electron-coupled to the *H105* output terminal through the left-hand section of the 6K8, pin terminal 3, *C122* and *R131.*

V110 is a reactance modulator, permitting frequency modulation of the 6K8 oscillator, which is required in the alignment of modern f-m and television sets and which is a very valuable feature. A signal voltage derived from *H108,* the EXT. F.M. terminal, is applied to *R147* and *R148.* A portion of this signal voltage is picked off by *R147* and applied to the grid circuit of *V110* through *R146* and *R144.* Capacitor *C137* is a compensating element in the network, designed to give linear response as the setting of *R147* is changed, during operation of the equipment. *C134* is an adjustment for getting proper f-m action. The plate-cathode circuit and input impedance of *V110* combine to form, in effect, a variable inductance shunt across the lower portion of *L02,* since the reactance of *C136* is very low. This inductive shunt reactance, when varied at a definite rate, changes the inductance of the tuned circuit, and, therefore, the operating frequency, thus providing the desired frequency modulation of the oscillator signal.

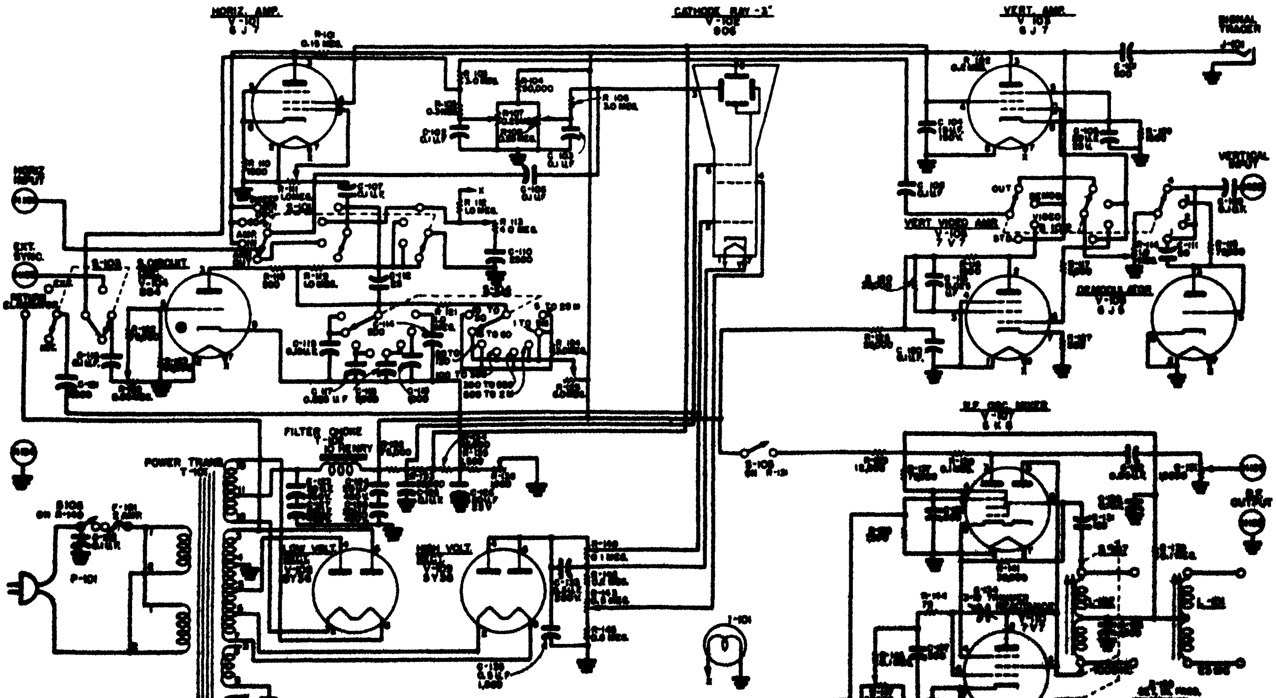


Fig. 22-31.—Schematic of Hickok Model RFO-5.

Courtesy Hickok Elec. Inst. Co.

HICKOK MODEL 195

Frequency Response
Vertical Amplifier 10 cps to 50 kc
Horizontal Amplifier 10 cps to 50 kc

Deflection Factors
Vertical-Deflection Plates 0.05 rms volts/inch
Vertical Amplifier 15 rms volts/inch
Horizontal Amplifier 0.3 rms volts/inch
Horizontal-Deflection Plates 30 rms volts/inch

Line Rating 105-125 volts, 50-70 cps

Tube Complement

Type	*Function*
6SJ7 (V1)	Horizontal Amplifier
884 (V2)	Sweep Circuit Oscillator
6AC7 (V3)	Vertical Amplifier
6X5GT (V4)	Low-Voltage Rectifier
5Y3GT (V5)	High-Voltage Rectifier
5UP1	Cathode-Ray Tube

The schematic circuit diagram of Model 195 is shown in Fig. 22-32. One feature of this otherwise conventional instrument is the use of a phasing control, *R3*, to apply a voltage derived from the *V4* plate circuit to *C8* through the *S1* horizontal selector switch. *C11*, *C9*, *R8*, *R4*, *C6*, and *R3* make up the phasing network. There is a voltage division of the 350-volt, 60-cycle signal, between *C11* and *C9*. The signal voltage across *C9* is applied to *R12* through *S1*, *C8*, and *R8*. *R3* affects the amplitude and phase by changing the circuit impedance.

HICKOK MODEL 195B

Frequency Response
Vertical Amplifier 30 cps to 1,000 kc
Horizontal Amplifier 10 cps to 50 kc
Sweep Circuit 10 cps to 25 kc

Deflection Factors
Vertical Amplifier 0.03 rms volts/inch
Vertical-Deflection Plates 15 rms volts/inch
Horizontal Amplifier 0.3 rms volts/inch
Horizontal-Deflection Plates 30 rms volts/inch

Line Rating 105-125 volts, 50-70 cps

Tube Complement

Type	*Function*
6SJ7 (V1)	Horizontal Amplifier
884 (V2)	Sweep Circuit Oscillator
6SN7 (V3)	Vertical Cathode-Follower Input Amplifier
6AC7 (V4)	Vertical Amplifier
6X5GT (V5)	Low-Voltage Rectifier
5Y3GT (V6)	High-Voltage Rectifier
5UP1	Cathode-Ray Tube

The schematic circuit diagram of Model 195B is shown in Fig. 22-33. This oscilloscope provides for external *Z*-axis or intensity grid modulation of the cathode-ray tube. A signal may be fed to pin 2 of the cathode-ray tube through *C31* and *S5*, using either external or internal modulation. The horizontal sweep and remaining circuits are very much like those of Hickok Model 195. However, the phasing control and filter circuits are a little different. Two filters, *C32* and *C33*, are used instead of four filters. The phasing voltage is derived from the transformer primary rather than from the high-voltage secondary.

HICKOK MODEL 305

Frequency Response
Vertical Amplifier 30 cps to 1 Mc
Horizontal Amplifier 10 cps to 50 kc
Sweep Circuit 10 cps to 25 kc
F.M. Oscillator 0-30 kc, and 0-450 kc

Line Rating 115 volts, 60 cps

The tube complement and tube functions of Model 305 are given in the schematic diagram, Fig. 22-34. This oscilloscope is similar to Hickok Model RFO-5. It should be noted, however, that Model RFO-5 uses 1,000 kc and 23 Mc for the r-f oscillator, while Model 305 uses 1,000 kc and 50 Mc. The use of the *V6* cathode-follower also permits the employment of a low-impedance vertical gain control *R50* of 10,000 ohms in place of the high-impedance control *R114* in Model RFO-5. The low-impedance control has less frequency discrimination and is, therefore, an improvement. The signal tracer jack *J1* is in the low-impedance cathode circuit of *V5*, rather than in the high *Z*-plate circuit of *V103* in the RFO-5. A connection to it, therefore, is a less critical matter. High-frequency compensation in the form of *L2* and *L3* is used to give a wide-band characteristic in the vertical-amplifier stage.

HICKOK MODEL 505A

Frequency Response
Vertical Amplifier 30 cps to 1 Mc
Horizontal Amplifier 10 cps to 50 kc
Sweep Circuit 10 cps to 25 kc

Deflection Factors
Vertical Amplifier 0.08 rms volts/inch
Vertical-Deflection Plates 15 rms volts/inch
Horizontal Amplifier 0.13 rms volts/inch
Horizontal-Deflection Plates 30 rms volts/inch

Line Rating 105-125 volts, 50-70 cps

The tube complement and its functions for Model 505A are shown in the schematic circuit diagram, Fig. 22-35. Model 505A is very similar to the 305 electrically.

JACKSON MODEL CRO-1

Frequency Response
Vertical Amplifier—Wideband 20 cps to 4½ Mc
Vertical Amplifier—High Sensitivity 20 cps to 100 kc
Horizontal Amplifier 20 cps to 150 kc
Sweep Circuit 20 cps to 50 kc

Deflection Factors
Vertical Amplifier—Wideband 0.25 rms volts/inch
Vertical Amplifier—High Sensitivity 0.018 rms volts/inch
Vertical-Deflection Plates 12 rms volts/inch
Horizontal Amplifier 0.55 rms volts/inch
Horizontal-Deflection Plates 15 rms volts/inch

Line Rating 100-125 volts, 50-60 cps

Tube Complement

Type	*Function*
5Y3 (V1, V2)	Rectifiers
6J6 (V3, V4, V5)	Amplifiers
6J6 (V6)	Sweep Oscillator
6C4 (V7, V8)	Cathode Followers
5UP1 (V9)	Cathode-Ray Tube

The schematic circuit diagram for Model CRO-1 is shown in Fig 22-36. The vertical amplifiers are operated in a high-gain circuit to provide a sensitivity of 0.018 volts per inch in the *high-sensitivity* position. They are switched to video-type amplifiers in the *wideband* position and provide 0.25 volts per inch sensitivity with a frequency response within 1.5 db to 4.5 Mc. Intensity modulation may be applied to the grid of the cathode-ray tube at 60 cps internally, or an external voltage may be applied to a binding post on the front panel.

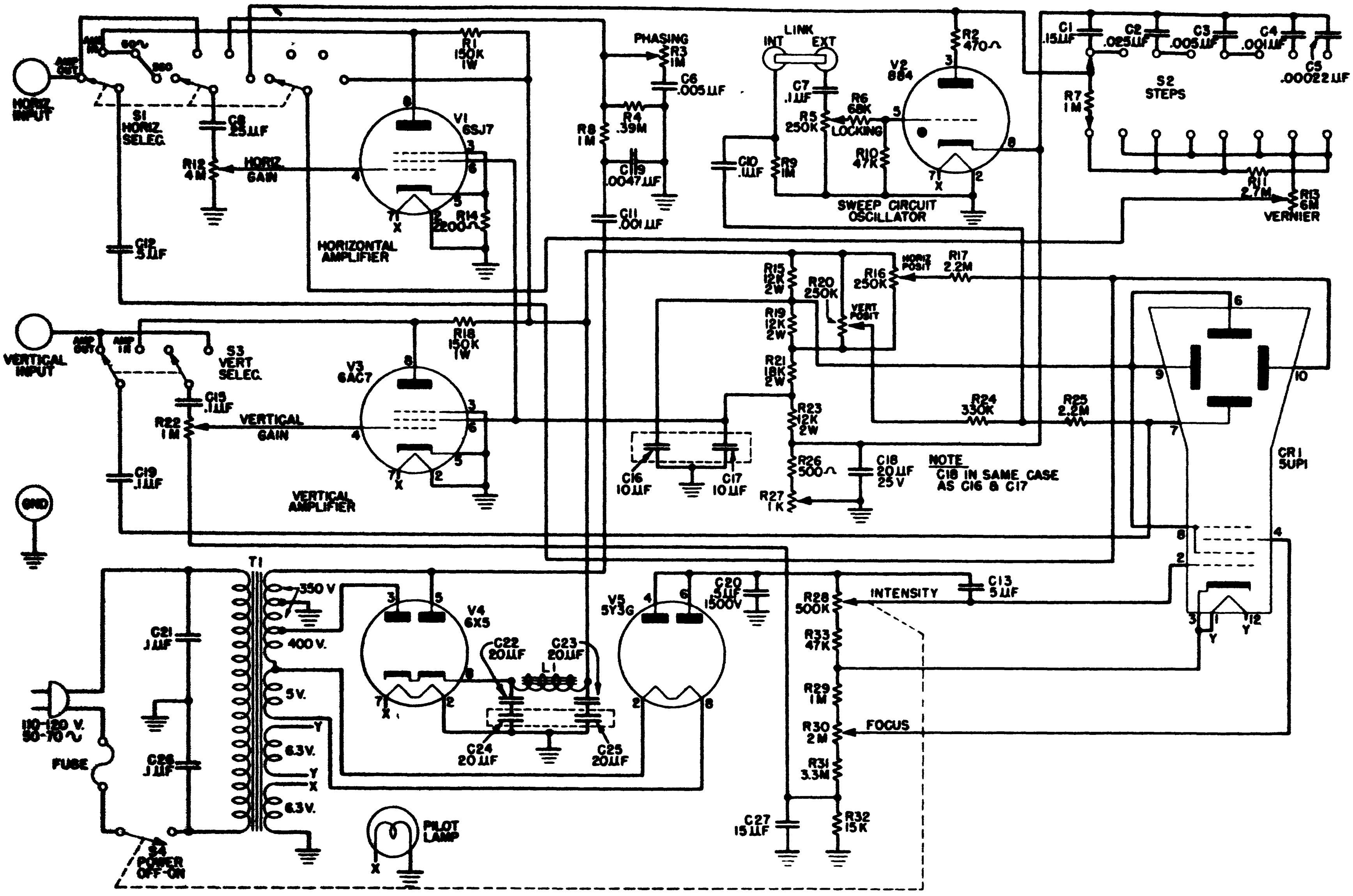


Fig. 22-32.—Schematic of Hickok Model 195.

Courtesy Hickok Elec. Inst. Co.

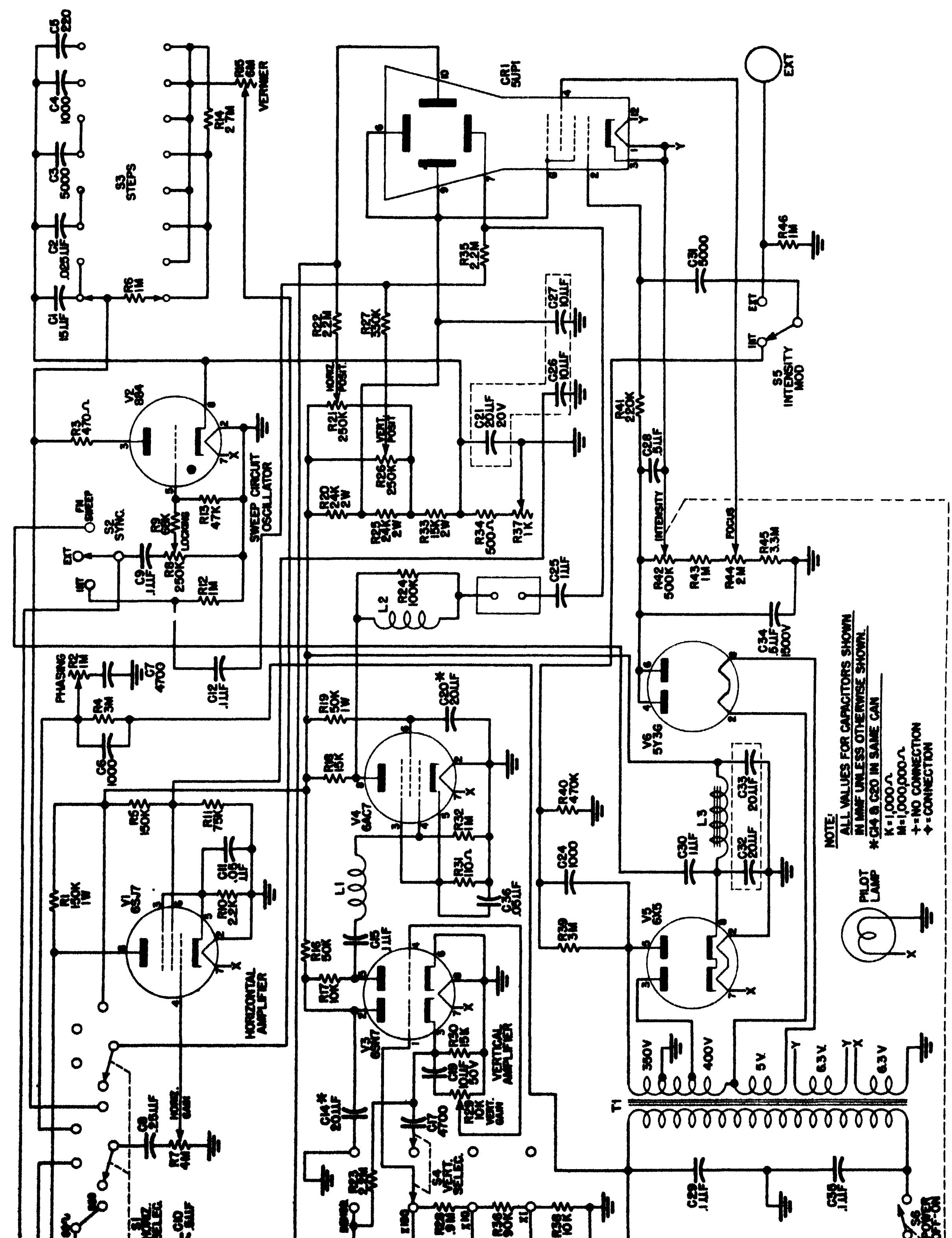


Fig. 22-33.—Schematic of Hickok Model 195-B.

Courtesy Hickok Elec. Inst Co.

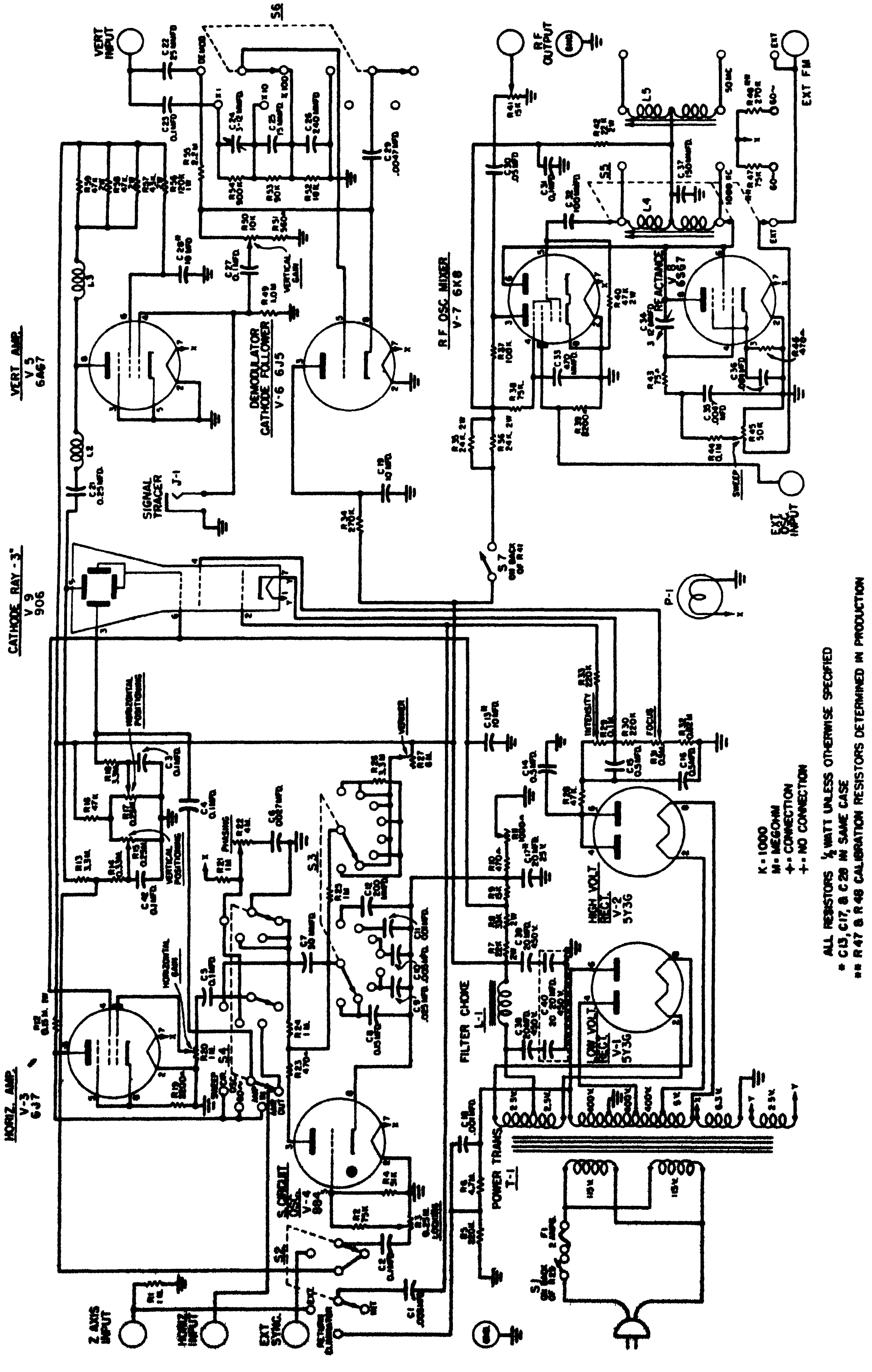


Fig. 22-34.—Schematic of Hickok Model 305.

Courtesy Hickok Elec. Inst. Co.

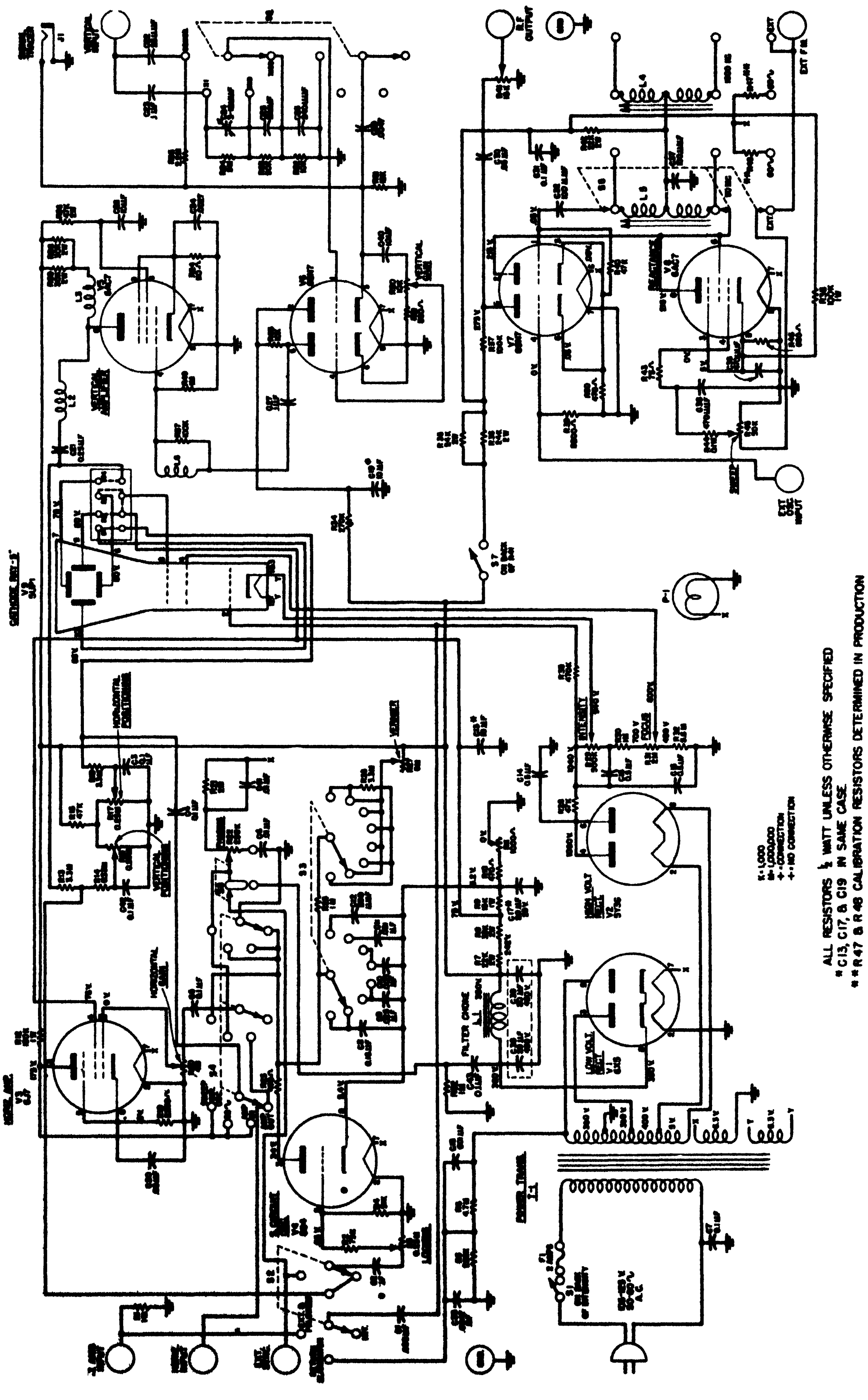
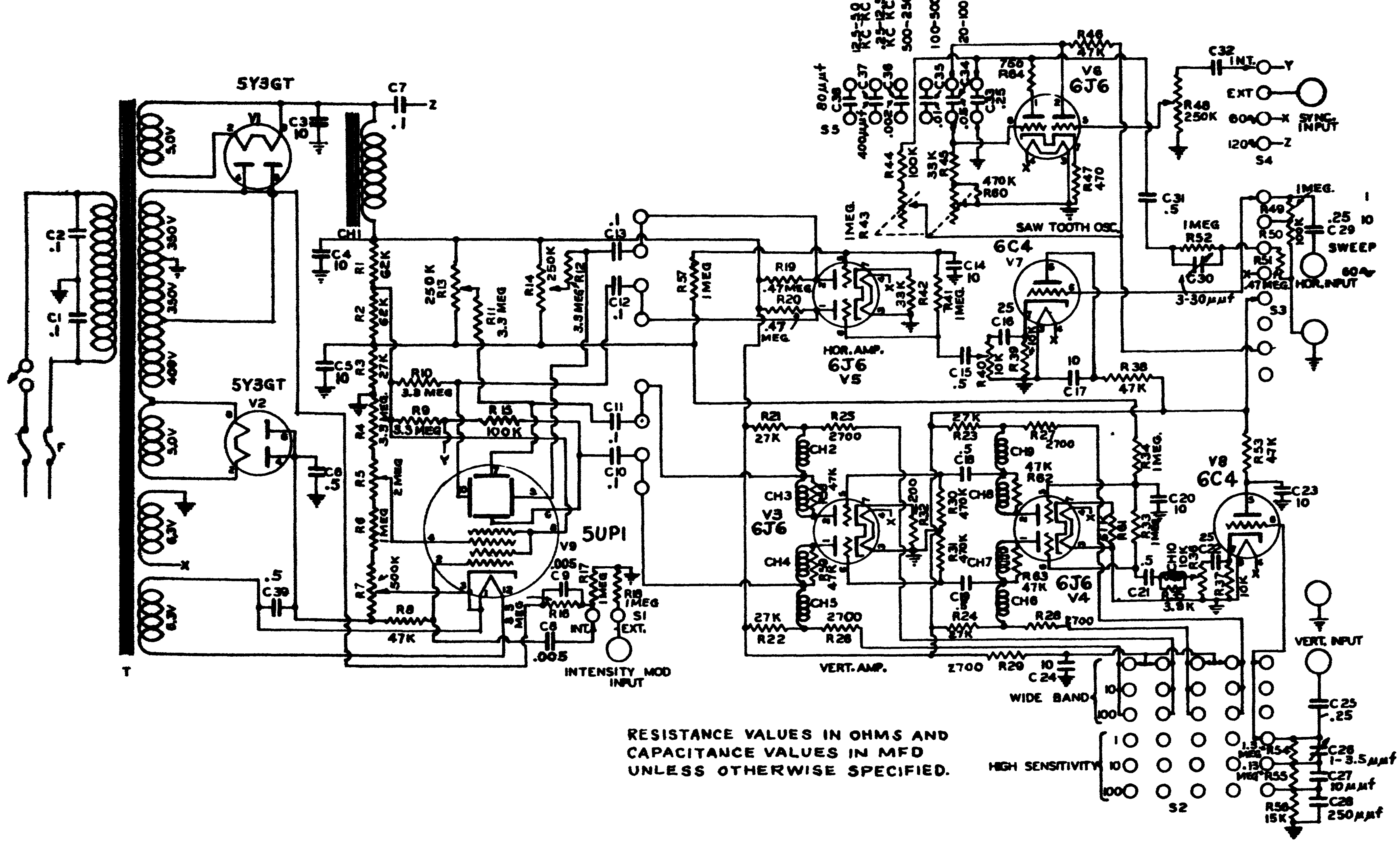


Fig. 22-35.—Schematic of Hickok Model 505A.

Courtesy Hickok Ele Inst. Co.



RESISTANCE VALUES IN OHMS AND CAPACITANCE VALUES IN MFD UNLESS OTHERWISE SPECIFIED.

Fig. 22-36.—Schematic of Jackson Model CRO-1

Courtesy Jackson Elec. Inst. Co.

MILLEN MODELS P4 AND P4-2

Synchroscope

LINE RATING 117 volts, 60 cps

TUBE COMPLEMENT

Type	*Function*
6SN7 (V1)	Sweep Generator
6SN7 (V2, V5)	Amplifier and Electronic Switch
6SN7 (V3, V4)	Trigger Amplifier
6SN7 (V6)	Amplifier and Trigger Control
6SN7 (V7)	Internal Free-Running Multivibrator
6SK7 (V8)	Vertical Discharge Tube
7V7 (V9)	Intensity Amplifier
6SN7 (V10)	Phase Splitter and Trigger Input Amplifier
2X2 (V11, V12)	High-Voltage Half-Wave Rectifiers
5Z3 (V13)	Low-Voltage Rectifier
5LP1 (V14)	Cathode-Ray Tube

The schematic circuit diagram for Models P4 and P4-2 are shown in Fig. 22-37. The synchroscope has an unusual sweep circuit which may be synchronized to an external positive or negative triggering pulse.

First, pin terminals 2 and 3 of *V2* can be visualized as the plate and cathode terminals of a diode electronic switch. Such a switch conducts a current only when the cathode is negative with respect to the plate. In effect, a generator with its negative terminal connected to pin terminal 3 and positive to ground causes an electron flow through *V2,* 3 to 2, under such conditions. This occurs when the junction of *C14* and *R14* become negative during one-half of the cycle of voltage across the secondary of *T2.*

As a result of the passage of an electron current through 3-2 of *V2,* with switch *S8* thrown to the *up* position on the diagram, the grid of *V3,* pin 1, becomes negative with respect to ground and its cathode. The resulting bias potential on the *V3* grid is capable of cutting off the *V3* plate current and *triggering* the signal or pulse.

During the next half-cycle of voltage across the secondary of *T2,* when the cathode of *V2* is positive with respect to ground and its plate, *V2* does not conduct and there is no bias potential applied to the grid of *V3,* so *V3* functions as an amplifier and transmits the signal to *V4* and the trigger output circuit. *V5,* 2-3, works in a similar manner in conjunction with *V5* terminals 6-4-5. The grid of *V5,* pin 4, goes directly to the plate, pin 2, in the other section of this dual-purpose tube.

During the time that the cathode of *V2,* pin 3, is positive with respect to ground, that of *V5,* pin 3, is negative or of opposite polarity. To change the polarity relationships or phases of the various circuits, *S8* is provided. For the case analyzed, *S8* was thrown to the up position. It is a double-pole, double-throw switch and may be thrown to the down position to change the polarity 180° as described.

V5 works into *V6* which is a cathode-follower output circuit used to drive the vertical plate circuit of the cathode-ray tube. If a positive pulse is applied to the grid of *V6,* no appreciable effect results unless *V6* previously had been cut off by a negative pulse. If this was the case and *C26* was allowed to discharge through the resistance of *V8,* the potential B+ potential is now applied to *C26* through *V6* to replenish the charge and the pulse signal is communicated to the vertical plate circuit of the cathode-ray tube through *C52* which couples the voltage across *V8.*

On the negative half-cycle of input voltage to the grid of *V6,* the tube is cut off and *C26* discharges through *V8.* Normally, *V8* has a resistance inversely proportional to plate voltage but because of the high value of plate resistance in this tube and the further effect of the cathode-circuit resistor which introduces a certain amount of degeneration and stabilizing action, the plate resistance effectively remains constant. *L4* and *C58* form a resonant plate circuit for the multivibrator, coupled through *C57.* The minimum plate-load resistance is set for the most part by *R56.*

V9 is an intensity amplifier tube which transfers a signal to the grid of the cathode-ray tube through *C41.* The input excitation is selected by switch *S3b. S2* is a switch which varies the gain of *V9* through selection of various values of input circuit resistances *R50, R51* and *R52.* A video amplifier attachment is available for use with the synchroscope and has a top frequency limit of about 5 Mc.

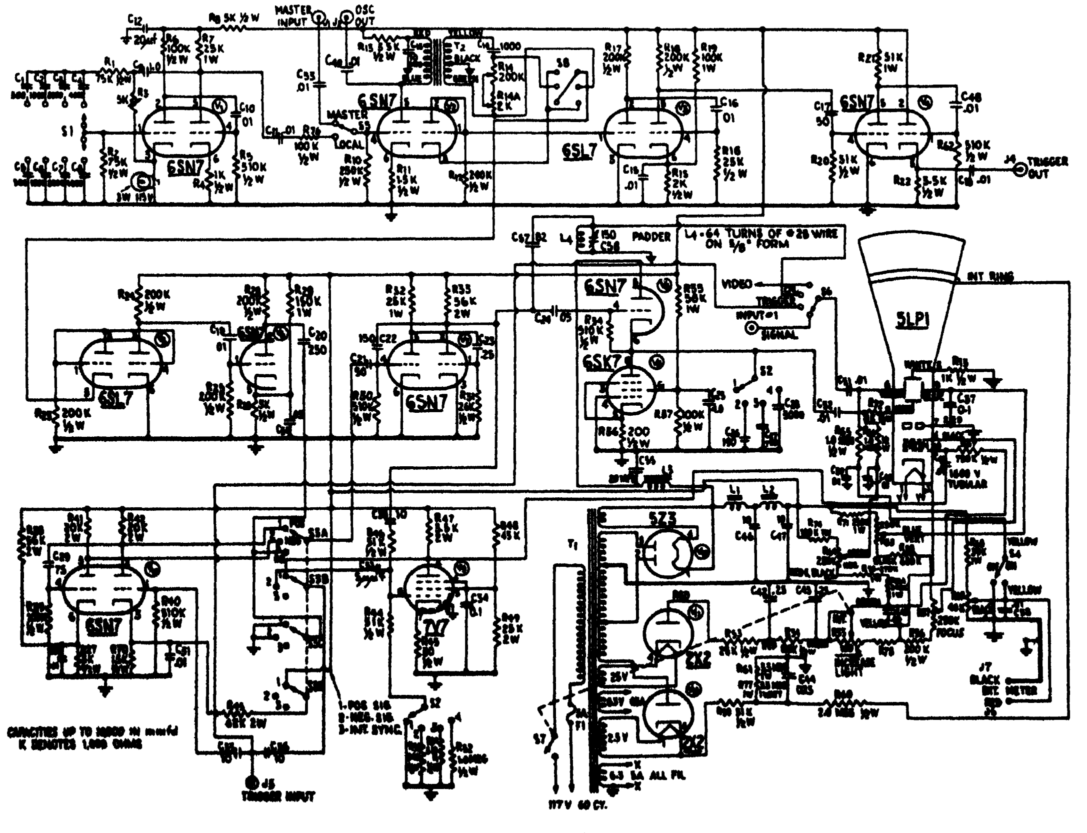


Fig. 22-37.—Schematic of Millen Models P4 and P4-2.

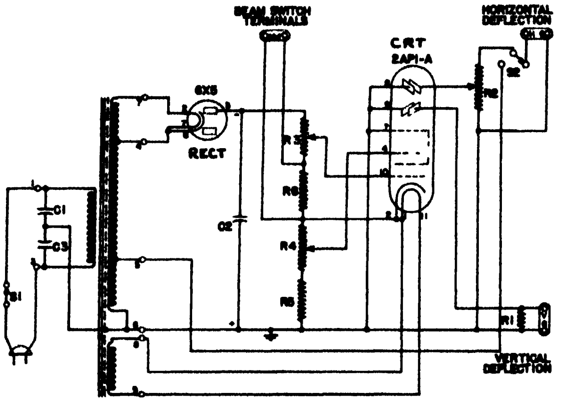
Courtesy Millen Mfg. Co.

NATIONAL MODEL CRU

Foundation Unit

Line Rating 115 volts, 60 cps

The schematic circuit diagram of Model CRU appears in Fig. 22-42. This instrument is a basic unit that may be used with additional apparatus. It contains only the cathode-ray tube and power supply.



C_1—.01 mfd. r.f. filter condenser
C_2—2. mfd. 600 volt (working) condenser
C_3—.01 mfd. r.f. filter condenser
R_1—4.7 megohm, ½ watt fixed resistor
R_2—500,000 ohm potentiometer (sweep voltage control)
R_3—50,000 ohm potentiometer (brilliancy control)
R_4—100,000 ohm potentiometer (focusing control)
R_5—220,000 ohm ½ watt fixed resistor
R_6—47,000 ohm ½ watt fixed resistor
S_1—Line switch
S_2—Horizontal sweep switch, s.p d.t.

Courtesy National Co.

Fig. 22-42.—Schematic of National Model CRU.

PANORAMIC MODEL PCA-2 TYPE T-200

Frequency Response
Vertical Amplifier R-F bandpass, 355 kc to 555 kc
R-F Sweep Circuit 581 kc to 781 kc
Line Rating 115 volts, 50-60 cps

Tube Complement

Type	*Function*
6SG7 (V1)	R-F Amplifier
6SA7 (V2)	Converter
6SG7 (V3)	I-F Amplifier
6SQ7 (V4)	Detector, Video Amplifier
6AC7 (V5)	Reactor
VR105 (V6)	Voltage Regulator
6SN7 (V7)	Sawtooth Generator and Amplifier
902A or 2AP1-A (V8)	Cathode-Ray Tube
6X5 (V9)	Low-Voltage Rectifier
6X5 (V10)	High-Voltage Rectifier

The schematic circuit diagram of the Panadaptor Model PCA-2 is shown in Fig. 22-43. This instrument is a special electronic development used for the visual observation of r-f signals over a band of frequencies that is relatively broad compared with the r-f band-pass characteristic of an ordinary radio set. *V1* is essentially a stage of r-f amplification of controllable gain. This gain is a function of the setting of *R3* in the cathode circuit which determines the bias voltage applied to the grid of *V1*. The signal output of *V1* is fed to the grid of *V2*, a 6SA7. *V2* is a mixer which combines the incoming signals and the locally generated oscillator signal provided by the oscillator section of *V2*.

V5 is a reactance-modulator tube used to frequency modulate the local oscillator of the Panadaptor. This frequency modulation action is dependent upon the amplitude of the signal voltage at grid pin 4 of *V5* and is derived from the cathode circuit of *V7*, pin 3, through *R14*, *R22*, *R23* and *C22*. The screen voltage of *V5* and that of *V2* are stabilized by the gaseous voltage regulator tube *V6*.

V3 is used in a typical i-f stage comparable to that found in a radio receiver, except that its selectivity is made as high as consistent with the sweep width and sweep rate. The detector is the diode section of a 6SQ7, *V4*. Its output is directly coupled to the grid of the triode section of the same tube so that the lowest possible signal frequency voltage, in this case d.c., appearing across the diode load resistor *R9*, will be amplified by the triode video amplifier.

During panoramic reception, the voltage pulses of i.f. fed to the detector appear (rectified and filtered) across the diode load resistor *R9*. These negative voltage pulses are fed back through the filter *R7-C7* to the control grid of the i-f amplifier. Strong signals produce high negative voltage pulses which reduce the gain of the i-f stage. In this way, the amplitude of strong signals is automatically decreased and it becomes possible to present simultaneously signals which differ considerably in strength. The time constant of the filter is sufficiently short so that a pulse of a strong signal does not reduce the gain for an adjacent weak signal.

Sawtooth Generator and Amplifier

The sawtooth voltage applied to the modulator *V5* is obtained from a blocking oscillator, which is an unusual application for this type of circuit. The operation of a blocking oscillator is discussed fully in Chapter 9.

By introducing the power line frequency into the grid input circuit, through divider resistors *R19* and *R21* (not present in units using a type 902A cathode-ray tube), the sawtooth is locked or synchronized to one-half the power line frequency. The sawtooth voltage developed across *C20* is applied and amplified by the other half of the 6SN7.

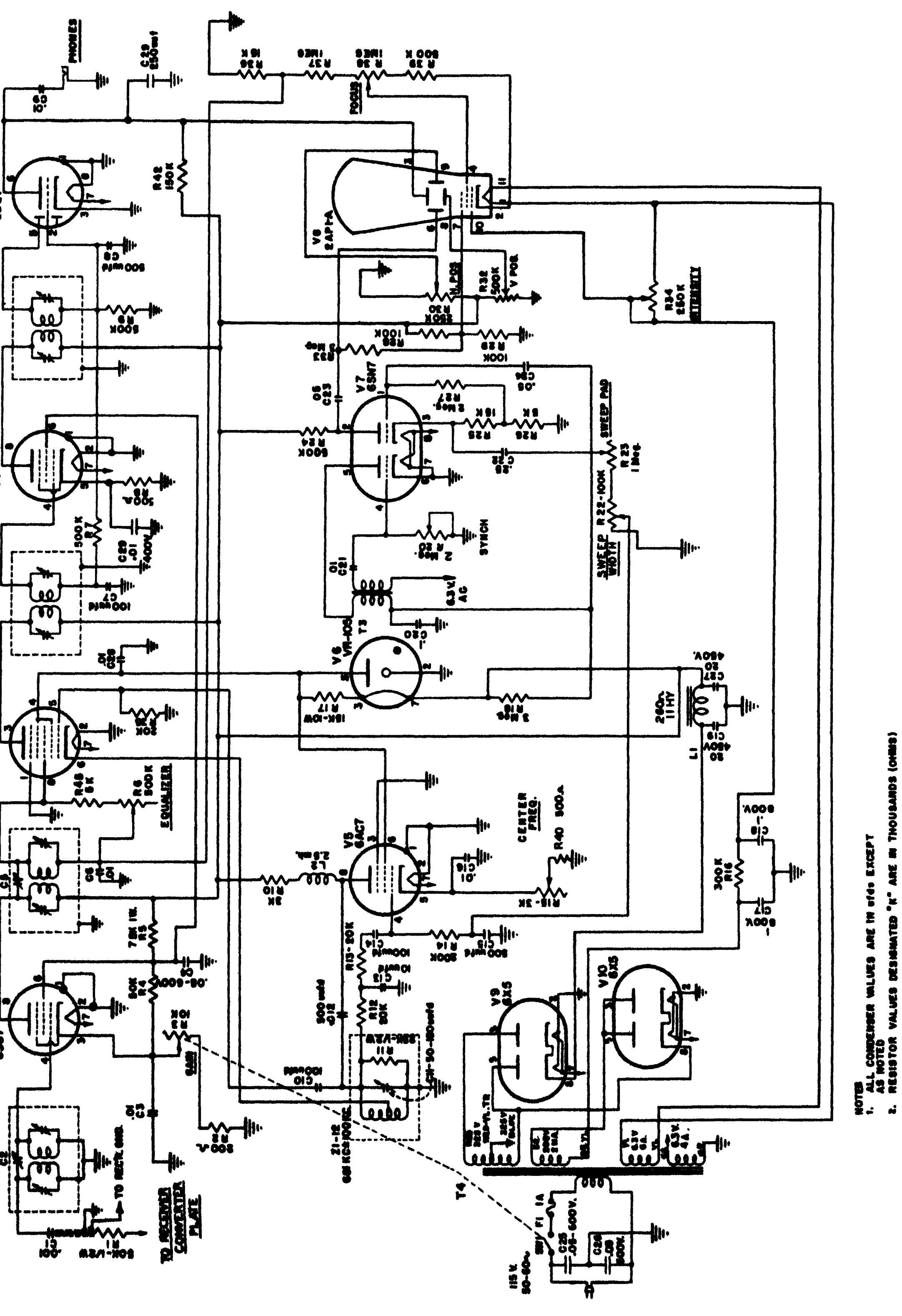
Another sawtooth output is taken at the cathode of the section and this is applied through potentiometer *R23*, the SWEEP PAD, and the SWEEP WIDTH control *R22* to the modulator tube. The function of the sweep pad is to proportion the proper amount of sawtooth signal across the sweep-width control so that a 200-kc oscillator excursion is obtained with the sweep-width control set to maximum. The fact that the same sawtooth voltage is used to swing both the electron beam and the oscillator frequency simultaneously is necessary for understanding the development of the Panoramic display.

Since a given instantaneous value of sawtooth voltage will correspond to one particular horizontal location on the tube face, and to one particular frequency of the oscillator, signals across the band being examined will be spread across the face of the cathode-ray tube in exactly the same manner that the signals would be spread across the range of a tuning dial.

Panoramic Operation

During operation, the oscillator is "rocked" through a band of frequencies by a linear-sawtooth voltage taken off the SWEEP-WIDTH control *R22*, and applied to the control grid of the modulator tube *V5*. The oscillator, thus, is frequency-modulated, and the extent of the deviation is proportional to the amplitude of the applied sawtooth voltage. The position of the SWEEP-WIDTH control contact arm determines the amount of sawtooth voltage applied to the modulator grid.

Linear-frequency calibrations on the Panoramic screen are obtained by having the oscillator frequency-excursion follow the sawtooth voltage linearly. A special phase net made up of *R12*, *R13*, and *C13* accomplishes this task. The design of the net is critical, but once it has been adjusted properly, it is highly stable. As the oscillator makes its excursion, it beats progressively and periodically with one incoming signal after another to produce pulses of i.f. at the 225-kc frequency, and these are passed and amplified by the i-f section.



NOTES
1. ALL CONDENSER VALUES ARE IN mfd. EXCEPT AS NOTED
2. RESISTOR VALUES DESIGNATED "K" ARE IN THOUSANDS (OHMS)
3. C19 AND C27 CAN BE 2 x 15 MFD-450 V.

Courtesy Panoramic Radio Corp.

Fig. 22-43.—Schematic of Panoramic Model PCA-2, type T-200.

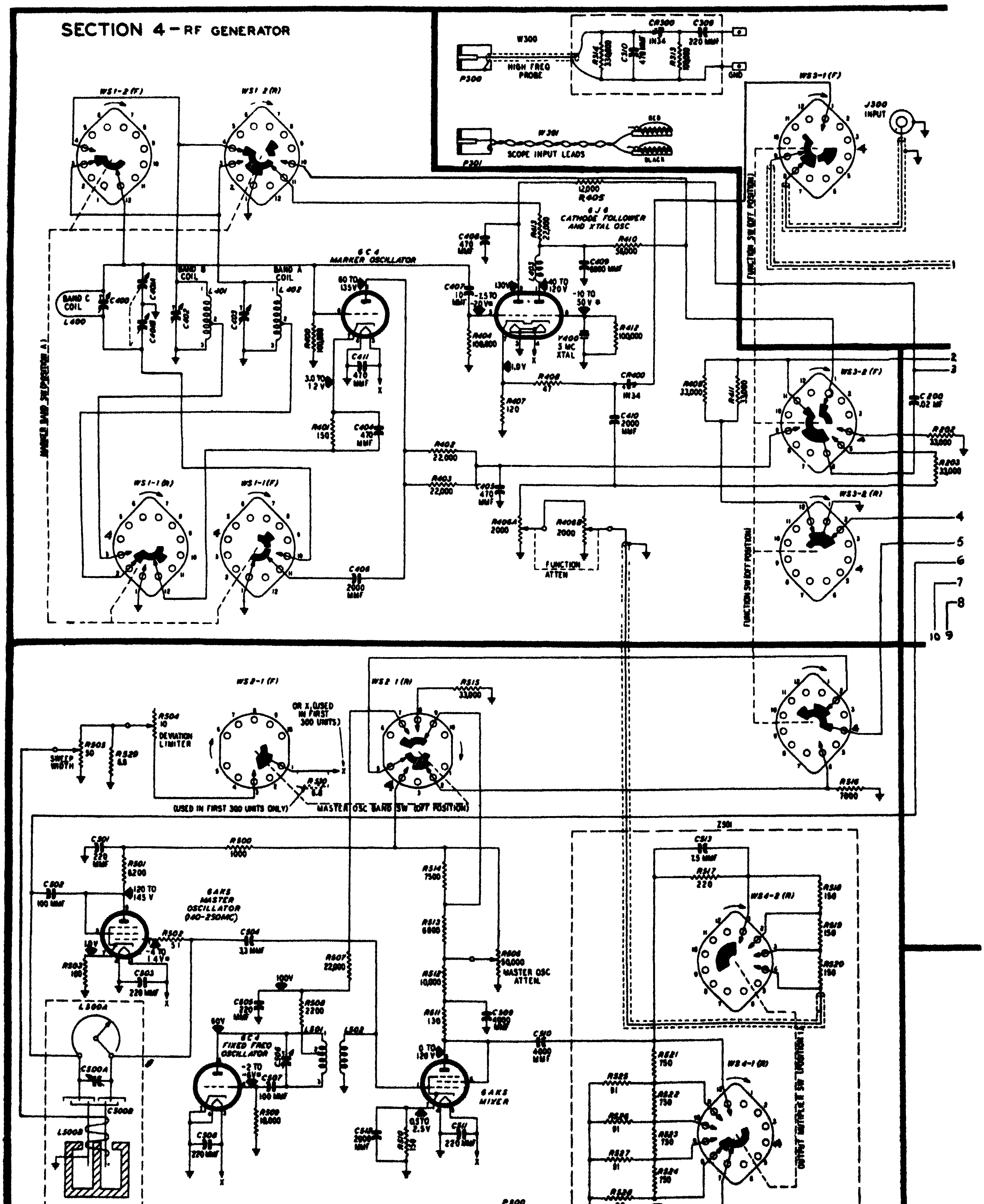
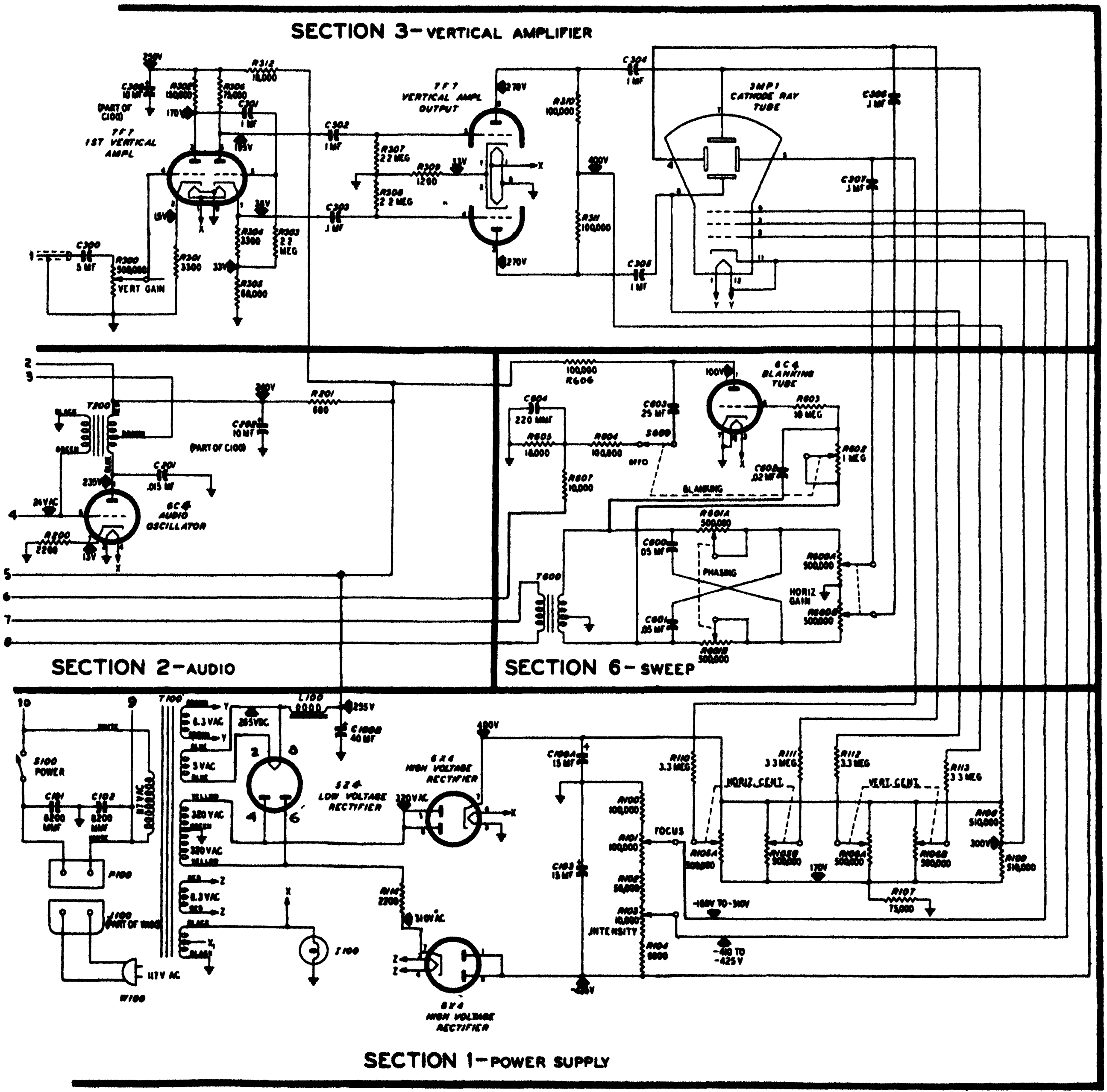


Fig. 22-44.—Schematic of Philco Model 7008.

Courtesy Philco Corp.



PHILCO MODEL 7008 (VISUAL ALIGNMENT GENERATOR)

Frequency Response

Sweep Circuit 60 cps, Sine Wave

Line Rating 110-120 volts, 60 cps

The schematic circuit diagram, including the tube complement and its functions, is shown in Fig. 22-44. This is not a conventional cathode-ray oscilloscope, but one designed primarily for use as an alignment indicator in f-m and television alignment work. A 60-cps internal sweep is used instead of the adjustable sweep found in the usual oscilloscope.

Blanking Circuit

The blanking circuit establishes a reference base line on the cathode-ray-tube screen, together with the response curve of the unit under test. It can be used only with the f-m-master oscillator generator. The baseline is produced by periodically removing the sweep signal applied to the output of the 7008 and allowing the horizontal sweep to trace a line across the screen during the absence of vertical deflection. This cycle of events occurs at a 60-cps rate, thus making it appear as if the response curve and baseline were produced simultaneously.

A 60-cps sine-wave voltage is applied to the grid of a 6C4 tube from the phasing transformer through a phase-shifting network consisting of a .02-μf capacitor *(C602)* and a 1-megohm blanking control *(R602)*. The flow of grid current through *R603*, a 10-megohm grid-current limiting resistor produces practically cut-off bias, so that no effect is obtained on the negative half-cycles. However, on the positive half-cycles, plate-current flow is increased, creating a negative swinging-gate effect in the plate circuit. This negative gate signal is applied through the on-off switch attached to the BLANKING control, and through a filter-and voltage-divider *R-C* network, to the grid of the 6AK5 master-oscillator tube.

When the negative gate signal appears at the grid of the 6AK5 oscillator tube, it prevents the tube from operating. Thus, the trace on the oscilloscope collapses to a straight line during the gating period. The alternate straight line and response-curve presentations appear as a response curve with a reference baseline on the cathode-ray-tube screen.

PRECISION MODEL ES-500

Frequency Response
Vertical Amplifier 10 cps to 1mc, within 6 db
Horizontal Amplifier 10 cps to 250 kc, within 1 db
Sweep Circuit 10 cps to 30 kc
Deflection Factors
Vertical Amplifier 0.02 rms volts/inch, maximum
Horizontal Amplifier 0.5 rms volts/inch
Line Rating 110-120 volts, 50-60 cps

Tube Complement

Type	*Function*
6J5	Vertical Input
6AK5	Vertical Amplifier
7W7	Vertical Amplifier
6SN7GT	Sweep Oscillator
7W7	Horizontal Amplifier
5Y3GT	Low-Voltage Rectifier
2X2	High-Voltage Rectifier
5CP1A	Cathode-Ray Tube

The schematic circuit diagram of Model ES-500 is shown in Fig. 22-45.

RADIO CITY PRODUCTS MODEL TV 90

Frequency Response
Vertical Amplifier 5 cps to 200 kc, ±2 db
Horizontal Amplifier 5 cps to 200 kc, ±2 db
Sweep Circuit 10 cps to 45 kc
Deflection Factors
Vertical Amplifier 0.285 rms volts/inch
Horizontal Amplifier 0.32 rms volts/inch
Line Rating 115 volts, 50-60 cps

Tube Complement

Type	*Function*
7G7 (V1)	Vertical Amplifier
7G7 (V2)	Horizontal Amplifier
3BP1 (V3)	Cathode-Ray Tube
6X5GT (V4)	High-Voltage Rectifier
6X5GT (V5)	Low-Voltage Rectifier
884 (V6)	Sweep Oscillator
7A4 (V7)	F-M Oscillator
6SB7Y (V8)	R-F Oscillator

The schematic circuit diagram for Model TV 90 is shown in Fig. 22-46. This instrument, primarily designed for television alignment, contains also an r-f oscillator and reactance modulator, besides the basic oscilloscope section There is also available a "travel" probe which may be used to pick up an r-f signal, demodulate it, and feed the signal to the vertical (Y) input of *V1*.

L4 in the *V7* f-m oscillator circuit is a special variable-inductance component which goes through a 60-cps modulation action due to the voltage at *R45*. *V7* is an ultra-audio type oscillator. Its f-m signal is fed into *V8* through *C25*. *V8* is also an oscillator, *C29*, *C28*, *L5* and the generated signals are mixed in *V8* to provide a high-frequency f-m output at *R52*. A marked generator can be connected to the marker jack.

RCA MODEL WO-27A

Frequency Response
Vertical Amplifier 0 cps to 100 kc
Horizontal Amplifier 0 cps to 100 kc
Sweep Circuit 30 cps to 100 kc
Deflection Factors
Vertical Amplifier 0.084 d-c volts/inch
Vertical-Deflection Plates Peak to peak 54 volts/inch
Horizontal Amplifier 0.105 volts/inch
Horizontal-Deflection Plates Peak to peak 67.5 volts/inch
Line Rating 105-115, 115-125 volts, 50/60 cps

Tube Complement

Type	*Function*
6SN7-GT (V1, V2, V4, V5)	Synchronizing Amplifier, Keying, Rectifier, Blanking Amplifier
6N7 (V3)	Timing-Axis Oscillator
6SF5 (V6, V7, V9, V10, V11, V12, V14, V15, V17, V18, V19, V20)	Voltage Amplifiers
5UP1 (V13)	Cathode-Ray Tube
OD3/VR-150 (V21, V22)	Voltage Regulators
5V4G (V23)	Low-Voltage Rectifier
6X5GT (V24, V25)	Voltage Doubler Rectifier
2X2A (V26)	High-Voltage Rectifier

The schematic circuit diagram of Model WO-27A is shown in Fig 22-47. The circuits are practically identical with those of Model 327-A, where a circuit analysis is given. The WO-27A uses a 5-inch rather than a 9-inch cathode-ray tube and a 2X2A rectifier rather than the 879 of the 327-A. The panel controls on the WO-27A and the general appearance of the instrument differ slightly from the 327-A. Other differences are negligible.

RCA MODEL WO-55A

Frequency Response
Vertical Amplifier 7 cps to 40 kc, flat within ± 10%
Horizontal Amplifier 7 cps to 40 kc, flat within ± 10%
Sweep Circuit 15 cps to 50 kc
Deflection Factors
Vertical Amplifier peak to peak, 1.33 volts/inch
Vertical-Deflection Plates peak to peak, 120 volts/inch
Horizontal Amplifier peak to peak, 1.5 volts/inch
Horizontal-Deflection Plates peak to peak, 135 volts/inch
Line Rating 105-125 volts, 50-60 cps

Tube Complement

Type	*Function*
6AU6 (V1, V2)	Horizontal Amplifiers
6AU6 (V3, V4)	Vertical Amplifiers
6J6 (V5)	Sweep Oscillator
6X4 (V6)	Low-Voltage Rectifier
6X4 (V7)	High-Voltage Rectifier
3MP1 (V8)	Cathode-Ray Tube

The schematic circuit diagram of Model WO-55A is shown in Fig. 22-48 The vertical and horizontal amplifiers use a standard push-pull and phase inverter circuit *V3* and *V2* functioning as grounded-grid amplifiers. However, they are somewhat unusual with respect to the vertical- and horizontal-centering circuit arrangements. Examination of the *V3*-*V4* section shows the electron flow from B—and ground is up through *R11*, *R10* and *R13* to pin 7 of *V3* and *V4* and from *R11* through *R8* and *R9* to the common cathode circuit, *R13* The bias on the *V4* grid is always one-half of that across *R10*, since *R8* and *R9* are equal. The bias on *V3* can be made any fraction of the voltage across *R10*. A minimum bias potential for both tubes, *V3* and *V4*, is set by *R13*. In practice, *R10* is adjusted visually for vertical centering by watching the beam spot on the cathode-ray-tube screen, increasing or decreasing the bias on *V3*, as required. *V1* and *V2*, in conjunction with *R23*, afford similar action for the horizontal amplifier. The sweep circuit is a Potter oscillator using a high-vacuum 6J6 twin triode *V5*.

RCA MODEL WO-58A

Frequency Response
Vertical Amplifier 5 cps to Mc, flat within ± 20%
Horizontal Amplifier 6 cps to 100 kc, flat within ± 10%
Sweep Circuit 10 cps to 100 kc
Deflection Factors
Vertical Amplifier through Attenuating Probe peak to peak, 5 volts/inch
Vertical Amplifier, Direct Input Probe peak to peak, 0.5 volt/inch
Vertical-Deflection Plates peak to peak, 39 volts/inch
Horizontal Amplifier peak to peak, 2 volts/inch
Line Rating 105-125 volts, 50-60 cps

The schematic circuit diagram for Model WO-58A is shown in Fig. 22-49. This model has been designed specifically for television servicing, and has some features that set it apart from the general run of oscilloscopes. First, three separate probes for connection to the vertical-amplifier input are provided: direct, attenuating, and crystal. The direct probe is used in low-level television stages where the shunting effect will not be excessive. Decreased coupling between the oscilloscope input and a television circuit is obtained with the attenuating probe, but there is some loss of signal due to the attenuation characteristic, The crystal probe is used for rectification action. The attenuating probe is useful for video-amplifier signal tracing in high levels because of its low input capacitance.

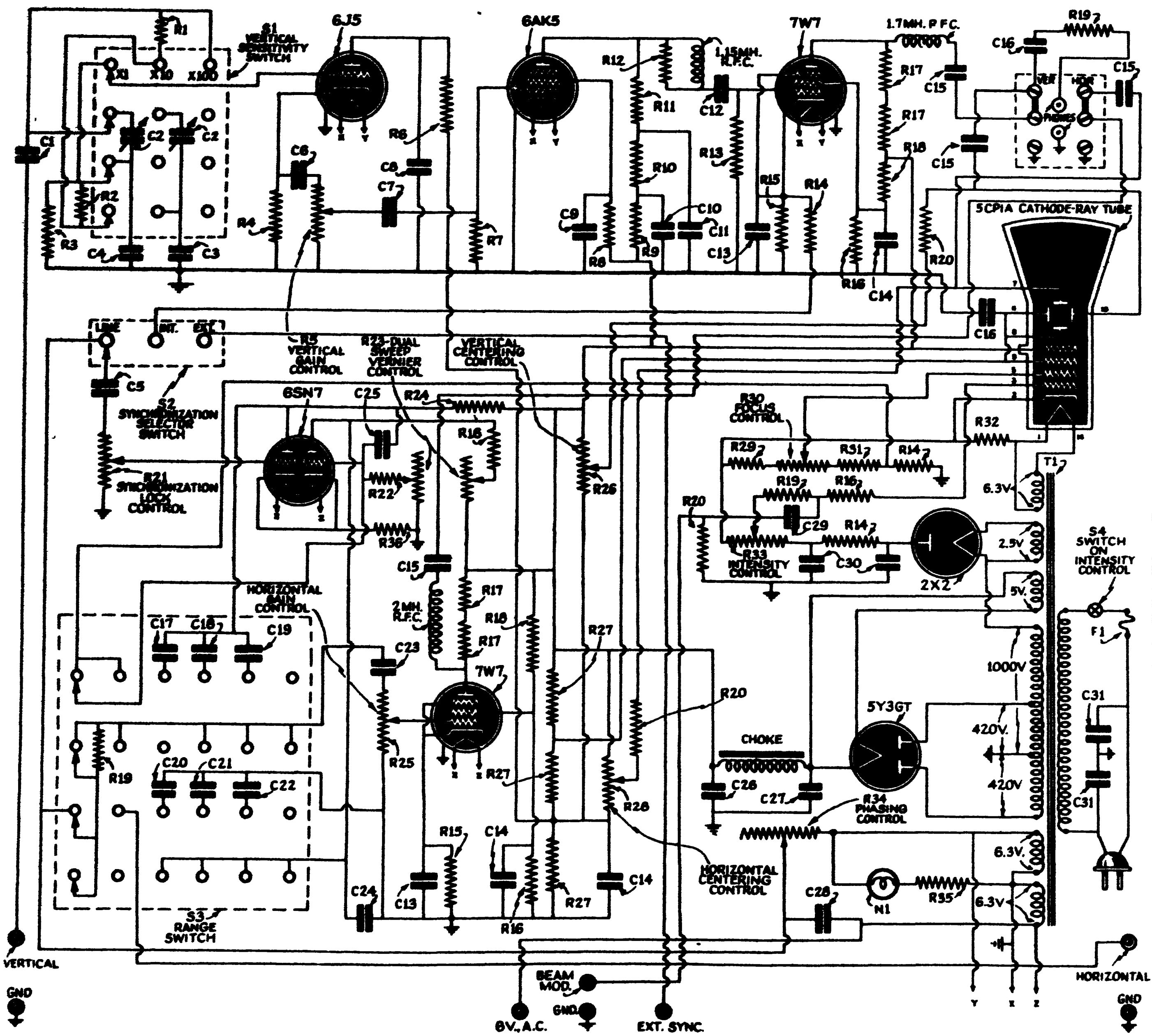


Fig. 22-45.—Schematic of Precision Model ES-500.

ITEM	PART NO.	SPECIFICATION
74	R36	200 Ω
73	R35	10 Ω
72	R34	500K Ω PHASING CONTROL
71	R33	100K Ω INTENSITY CONTROL
70	R32	10 MEG. Ω
69	R31	680K Ω
68	R30	500K Ω FOCUS CONTROL
67	R29	220K Ω
66	R28	500K Ω HOR CENTERING CONT.
65	R27	22K Ω EACH
64	R26	500K Ω VERT. CENTERING CONT
63	R25	0.5 MEG Ω HOR GAIN CONTROL
62	R24	30K Ω
61	R23	DUAL 500K Ω SWEEP VERNIER
60	R22	33K Ω
59	R21	10K Ω SYNC. LOCK CONTROL
58	R20	3.3 MEG Ω EACH
57	R19	1 MEG. Ω EACH
56	R18	56K Ω EACH
55	R17	10K Ω EACH
54	R16	100K Ω EACH
53	R15	160 Ω EACH
52	R14	56K Ω EACH
51	R13	560K Ω
50	R12	10K Ω
49	R11	10K Ω
48	R10	15K Ω
47	R9	5600 Ω
46	R8	120K Ω
45	R7	2.2 MEG. Ω
44	R6	1500 Ω
43	R5	10K Ω VERT GAIN CONTROL
42	R4	10K Ω
41	R3	20K Ω
40	R2	180K Ω
39	R1	1.8 MEG Ω
38	C31	.02-MFD. EACH
37	C30	0.5-MFD EACH
36	C29	.006-MFD.
35	C28	0.1-MFD.
34	C27	20-MFD.
33	C26	30-MFD.
32	C25	240 MMFD.
31	C24	.01-MFD.
30	C23	0.5-MFD
29	C22	.04-MFD.
28	C21	0.4-MFD.
27	C20	1 MFD.
26	C19	1250 MMFD.
25	C18	.012-MFD.
24	C17	0.11-MFD.
23	C16	.02-MFD. EACH
22	C15	0.25-MFD EACH
21	C14	40 MFD. EACH
20	C13	.0047-MFD EACH
19	C12	0.5-MFD
18	C11	4 MFD.
17	C10	12 MFD.
16	C9	15 MFD.
15	C8	15 MFD
14	C7	0.1-MFD
13	C6	25 MFD
12	C5	0.5-MFD
11	C4	1000 MMFD.
10	C3	82 MMFD.
9	C2	4-30 MMFD EACH
8	C1	0.25-MFD.
7	T1	POWER TRANSFORMER
6	S4	LINE SW. ON INTENSITY CONTROL
5	S3	RANGE SWITCH
4	S2	SYNC. SELECTOR SWITCH
3	S1	VERTICAL SENSITIVITY SWITCH
2	N1	#47 PILOT LAMP
1	F1	1 AMP. FUSE

Courtesy Precision Apparatus Co.

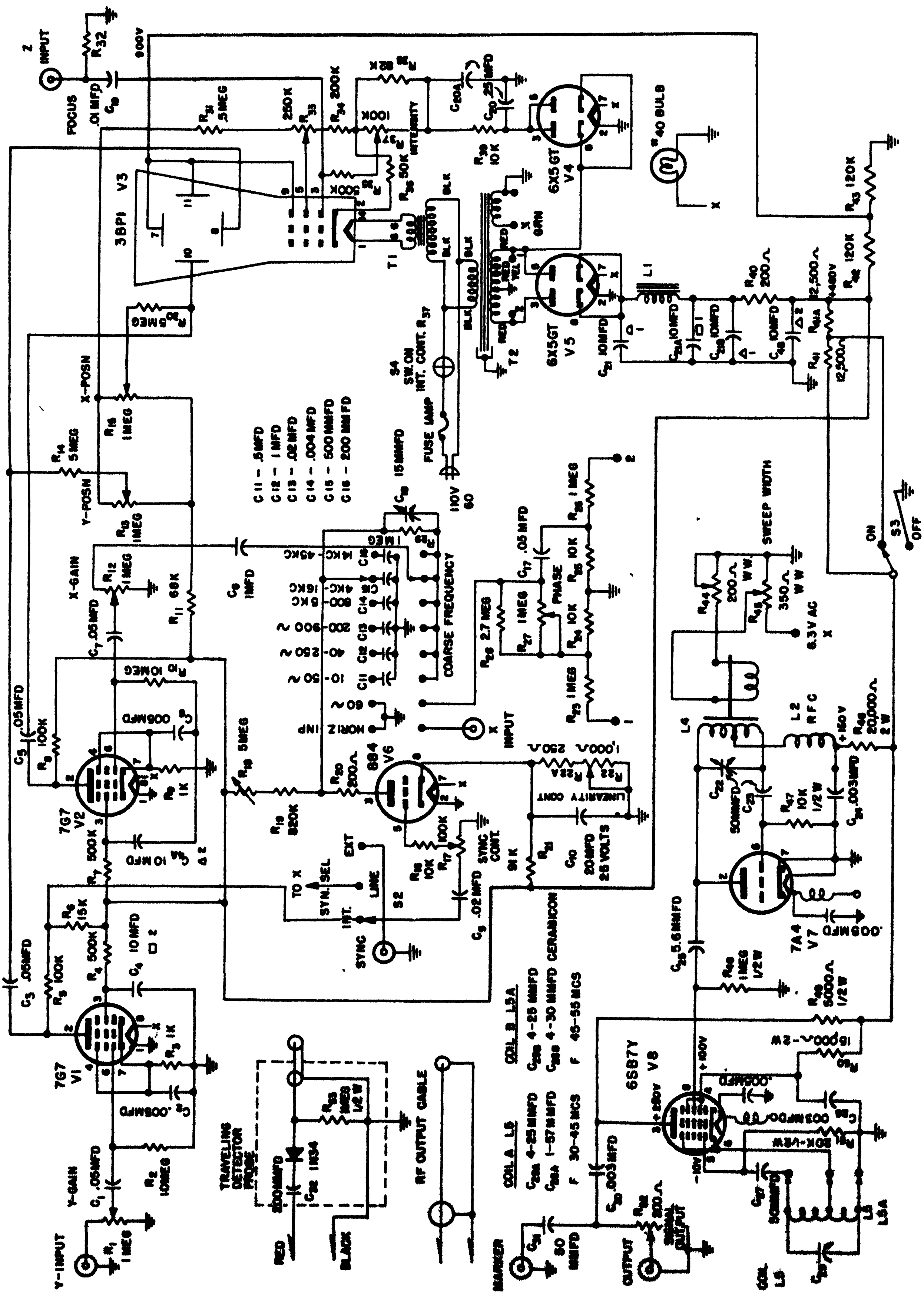


Fig. 22-46.—Schematic of Radio City Products Model TV-90.

Courtesy Radio City Prod. Co.

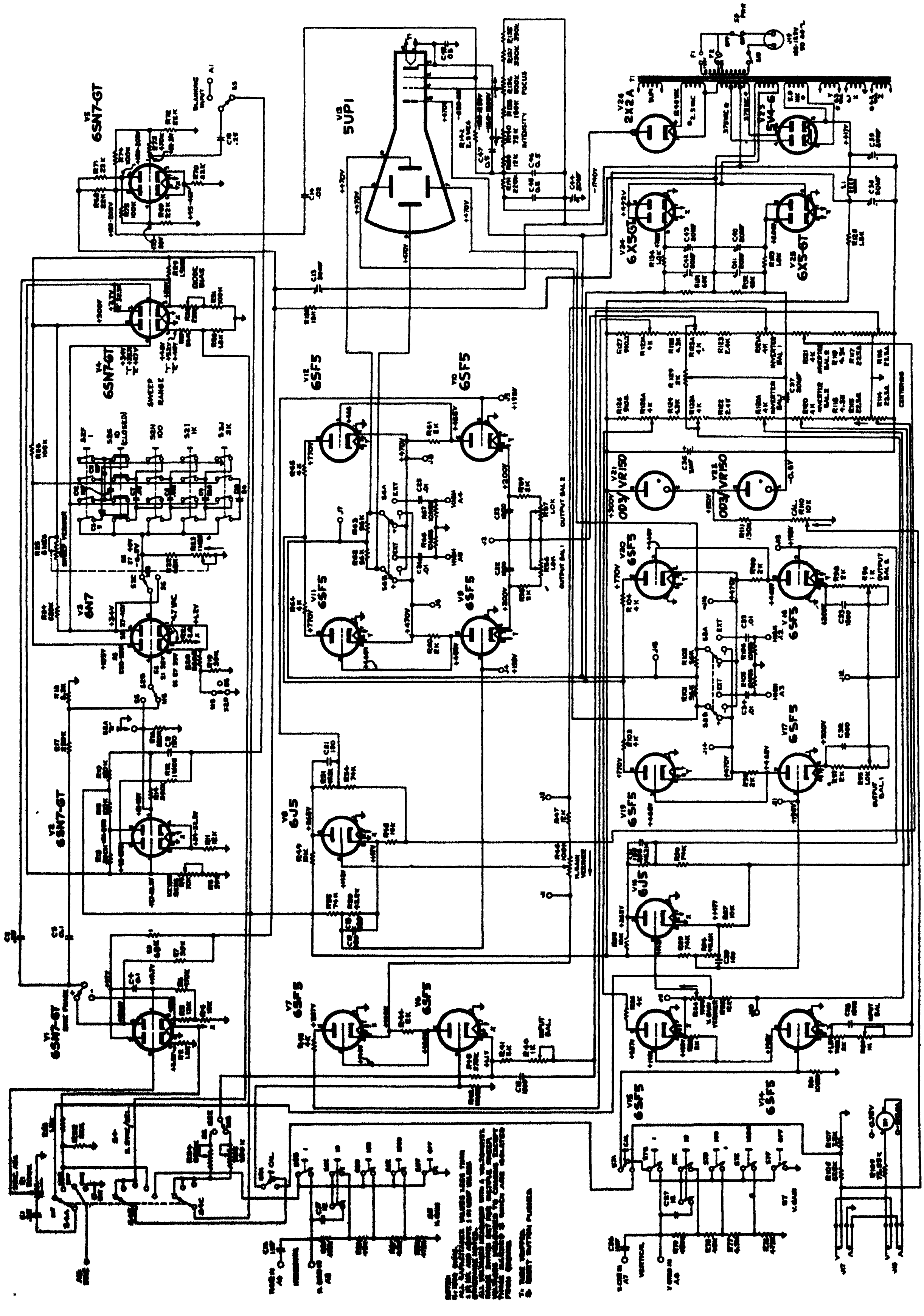
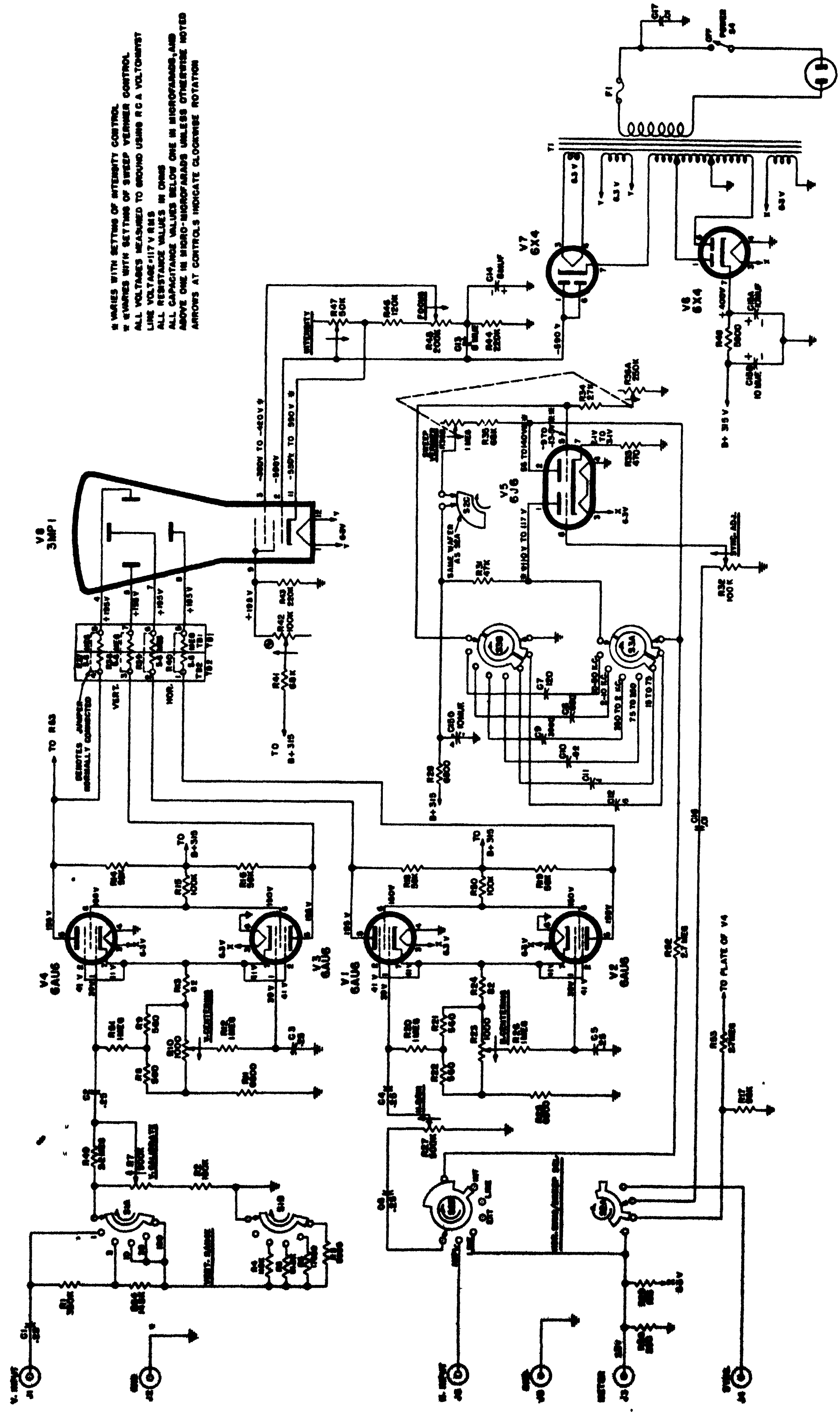


Fig. 22-47.—Schematic of RCA Model WO-27A.

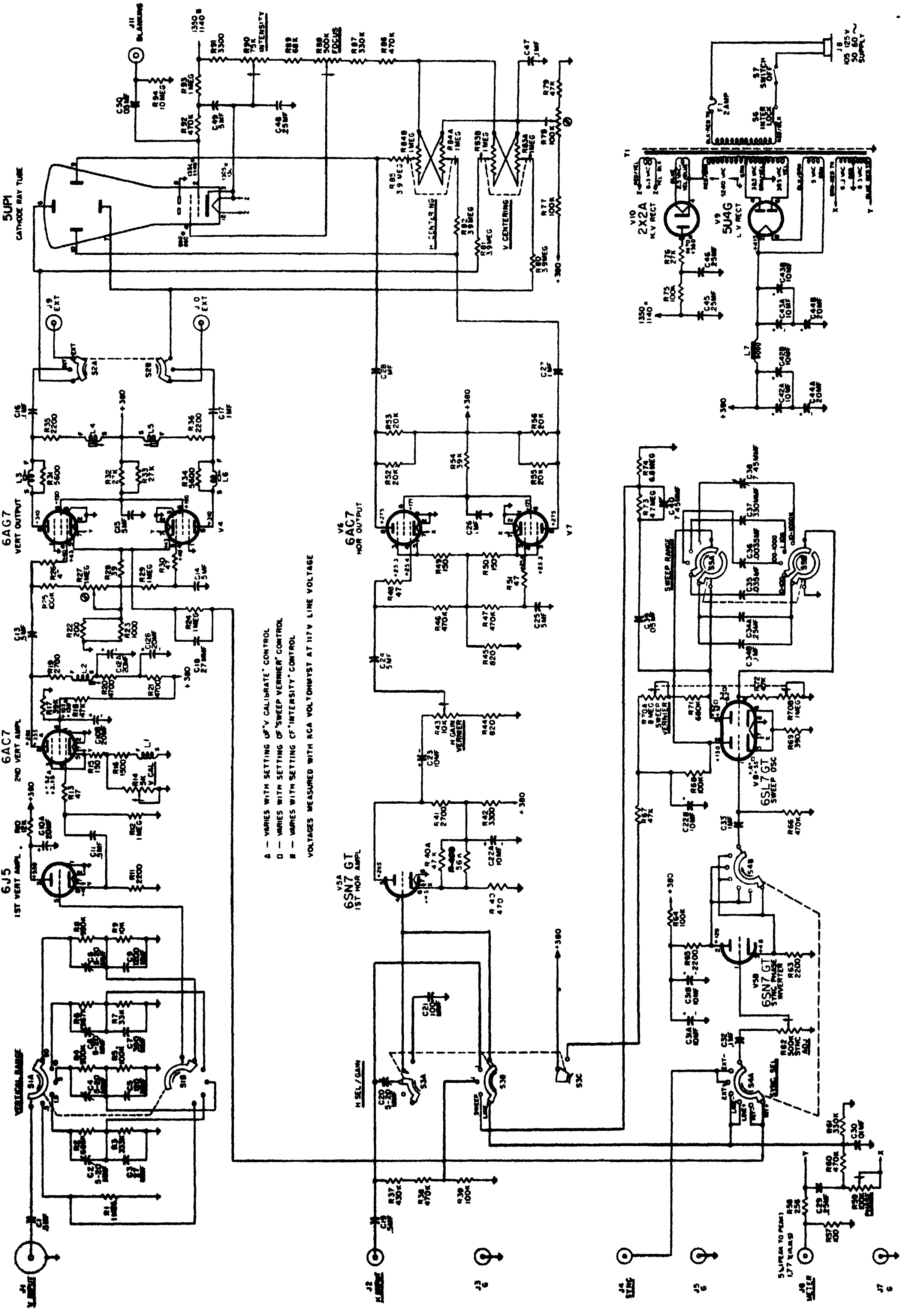
Courtesy RCA



* VARIES WITH SETTING OF INTENSITY CONTROL
** VARIES WITH SETTING OF SWEEP VERNIER CONTROL
ALL VOLTAGES MEASURED TO GROUND USING R C A VOLTOHMYST
LINE VOLTAGE=117 V RMS
ALL RESISTANCE VALUES IN OHMS
ALL CAPACITANCE VALUES BELOW ONE IN MICROFARADS, AND ABOVE ONE IN MICRO-MICROFARADS UNLESS OTHERWISE NOTED
ARROWS AT CONTROLS INDICATE CLOCKWISE ROTATION

Courtesy RCA

Fig. 22-48.—Schematic of RCA Model WO-55A.



Courtesy RCA

Fig. 22-49.—Schematic of RCA Model WO-58A.

RCA MODEL WO-60C

Frequency Response

Vertical Amplifier 2 cps to 100 kc, ± 20%
Horizontal Amplifier 2 cps to 100 kc, ± 20%
Sweep Circuit 3 cps to 30 kc

Deflection Factors

Vertical Amplifier 0.02 rms volts/inch
Vertical-Deflection Plates 12 rms volts/inch
Horizontal Amplifier 0.024 rms volts/inch
Horizontal-Deflection Plates 15 rms volts/inch

Line Rating 105-125 volts, 50-60 cps

The schematic circuit diagram of this conventional general-purpose instrument is shown in Fig. 22-50.

RCA MODEL WO-79A

Frequency Response

Vertical Amplifier 10 cps to 5 Mc, flat within ± 20%
Horizontal Amplifier 10 cps to 500 kc, flat within 10%
Sweep Circuit 20 cps to 250 kc, sawtooth 1 cps to 50 kc, triggered sweep

Deflection Factors

Vertical Amplifier peak to peak, 0.75 volts/inch
Vertical-Deflection Plates peak to peak, 68 volts/inch
Horizontal Amplifier peak to peak, 2 volts/inch
Horizontal-Deflection Plates peak to peak, 89 volts/inch

Line Rating 105-125 volts, 50-60 cps

The schematic circuit diagram of Model WO-79A, including tube complement and functions, is shown in Fig. 22-51. The vertical input is fed through a three-step attenuator, providing 0, 10 to 1, and 100 to 1 reduction, to the grid of the first amplifier *V1*, a 6AC7, which operates as a single-ended stage. Output taken from the cathode is fed through a time-delay network and voltage-dividing potentiometer to the grid of the second vertical amplifier. The time-delay network delays the vertical-deflection voltage two-tenths of one microsecond so that when the signal under observation is used to trigger the TIME OSC., the sweep will normally start before the signal reaches the deflection plates of the cathode-ray tube. The time delay thus facilitates observation of signals having steep wavefronts.

The instrument provides sawtooth as well as triggered sweep operation. An intensifying amplifier permits increasing the brilliance of the trace, when used for triggered sweep operation. An astigmatism control is provided consisting of a 100,000-ohm potentiometer *R93*, connected in the centering and focusing circuit. This control permits adjustment of deflection-plate potential with respect to the second anode of the cathode-ray tube for uniform definition over the entire surface of the screen.

RCA MODEL 151, 151A, 151-2

Frequency Response

Vertical Amplifier 20 cps to 15,000 cps
Horizontal Amplifier 20 cps to 15,000 cps
Sweep Circuit 30 cps to 10,000 cps

Deflection Factors (151, 151A)

Vertical Amplifier peak to peak, 5 volts/inch
Vertical-Deflection Plates peak to peak, 250 volts/inch
Horizontal Amplifier peak to peak, 5 volts/inch
Horizontal-Deflection Plates peak to peak, 250 volts/inch

Deflection Factors (151-2)

Vertical Amplifier 0.5 rms volts/inch
Vertical-Deflection Plates 30 rms volts/inch
Horizontal Amplifier 0.5 rms volts/inch
Horizontal-Deflection Plates 30 rms volts/inch

Line Rating 110-120 volts, 50-60 cps

Tube Complement

Type	*Function*
6C6	Vertical Amplifier
6C6	Horizontal Amplifier
885	Gaseous-Sweep Oscillator
913 (902)	Cathode-Ray Tube (1 inch)
80	Full-Wave Rectifier

The schematic circuit diagram of Model 151, 151A is shown in Fig. 22-52A and the difference between these models and Model 151-2 is shown in Fig. 22-52B.

There is one unique feature of this otherwise conventional circuit that causes unusual voltage distributions. Since the shell of the cathode-ray tube is connected to the second anode, which must be at a positive potential from the cathode, and since the shell must be grounded for safety, the positive side of the power supply must also be grounded. This is common practice in cathode-ray oscillopscopes, but in this case the power supply is common to the cathode-ray tube and the amplifier tubes. The cathode, grid, suppressor, and screen grids of the amplifiers are all at high potential to ground and the plate is nearly at ground potential.

While the voltage distribution is unusual, the method of operating the amplifier tubes has not been affected. The grids are maintained about two volts negative from the cathode, the suppressor is connected to the cathode, the screen grid is about 35 volts positive with respect to the cathode, and the plate is still more positive in each case.

RCA MODEL 155-A

Frequency Response

Vertical Amplifier to 12 kc, within 1 db
Horizontal Amplifier to 12 kc, within 1 db
Sweep Circuit 15 cps to 16 kc

Deflection Factors

Vertical Amplifier 0.8 rms volts/inch
Vertical-Deflection Plates 30 rms volts/inch
Horizontal Amplifier 0.8 rms volts/inch
Horizontal-Deflection Plates 30 rms volts/inch

Line Rating 110-120 volts, 50-60 cps

The schematic circuit diagram of Model 155-A is shown in Fig. 22-53.

RCA MODEL 155-C

Frequency Response

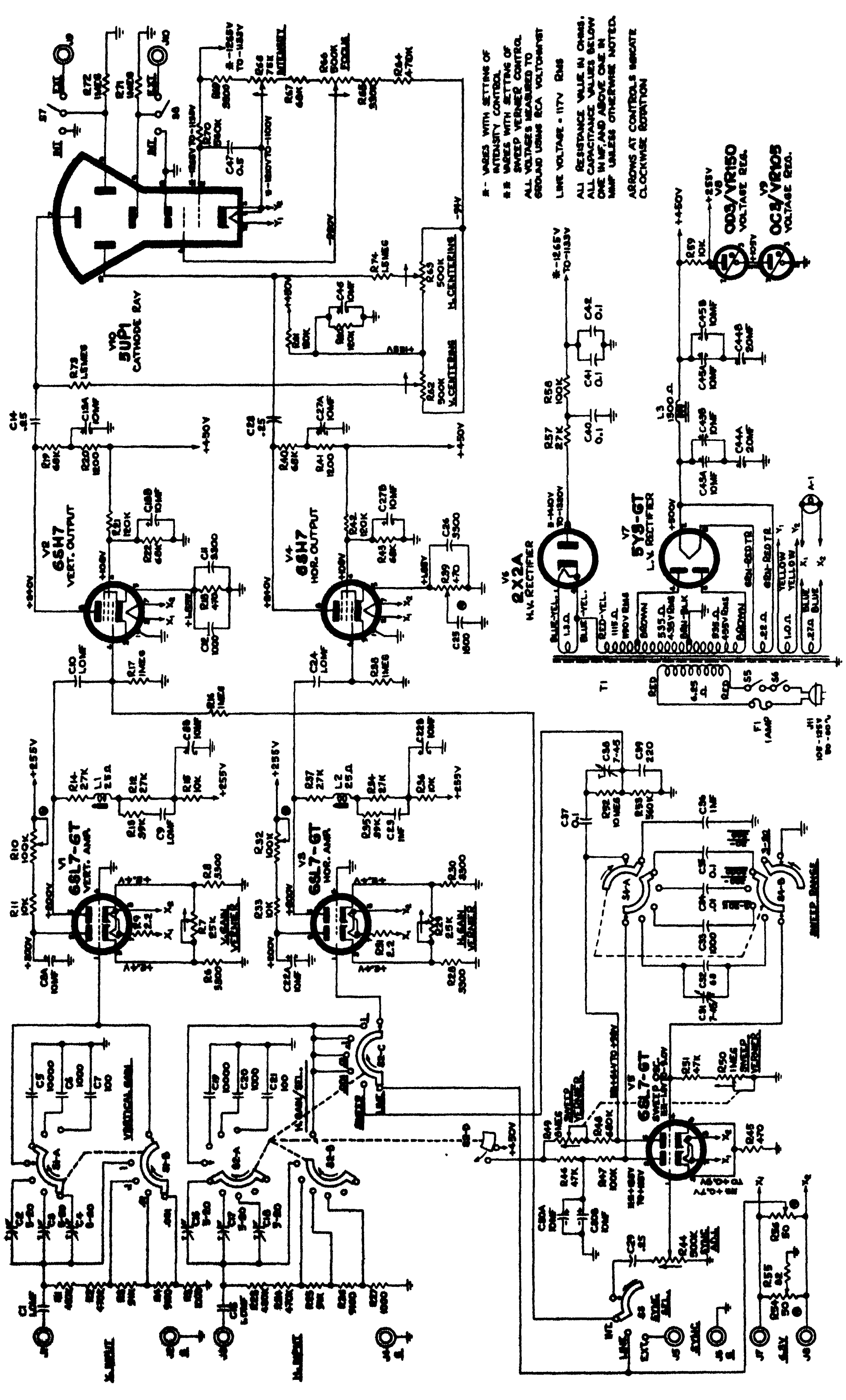
Vertical Amplifier 7 cps to 40 kc, ± 10%
Horizontal Amplifier 7 cps to 40 kc, ± 10%
Sweep Circuit 10 cps to 60 kc

Deflection Factors

Vertical Amplifier 1 rms volts/inch
Vertical-Deflection Plates 27 rms volts/inch
Horizontal Amplifier 1 rms volts/inch
Horizontal-Deflection Plates 27 rms volts/inch

Line Rating 110-120 volts, 50-60 cps

Model 155C uses a conventional circuit, as is apparent from Fig. 22-54. The 6C8G *(V3)* is a high-vacuum timing-axis oscillator tube used in a Potter oscillator circuit.


V1 6SL7-GT VERT. AMP.
V2 6SH7 VERT. OUTPUT
V3 6SL7-GT HOR. AMP.
V4 6SH7 HOR. OUTPUT
V5 6SL7-GT SWEEP OSC.
V6 2X2A H.V. RECTIFIER
V7 5Y3-GT L.V. RECTIFIER
V8 OD3/VR150 VOLTAGE REG.
V9 OC3/VR105 VOLTAGE REG.
V10 5UP1 CATHODE RAY
VERTICAL GAIN
H. GAIN/SEL.
V. GAIN VERNIER
H. GAIN VERNIER
SWEEP RANGE
SWEEP VERNIER
SYNC ADJ.
V. CENTERING
H. CENTERING
INTENSITY
FOCUS
T1
F1 1 AMP
* - VARIES WITH SETTING OF INTENSITY CONTROL
** VARIES WITH SETTING OF SWEEP VERNIER CONTROL
ALL VOLTAGES MEASURED TO GROUND USING RCA VOLTOHMYST
LINE VOLTAGE - 117V RMS
ALL RESISTANCE VALUE IN OHMS. ALL CAPACITANCE VALUES BELOW ONE IN MF, AND ABOVE ONE IN MMF UNLESS OTHERWISE NOTED.
ARROWS AT CONTROLS INDICATE CLOCKWISE ROTATION

Courtesy RCA

Fig. 22-50.—Schematic of RCA Model WO-60C.

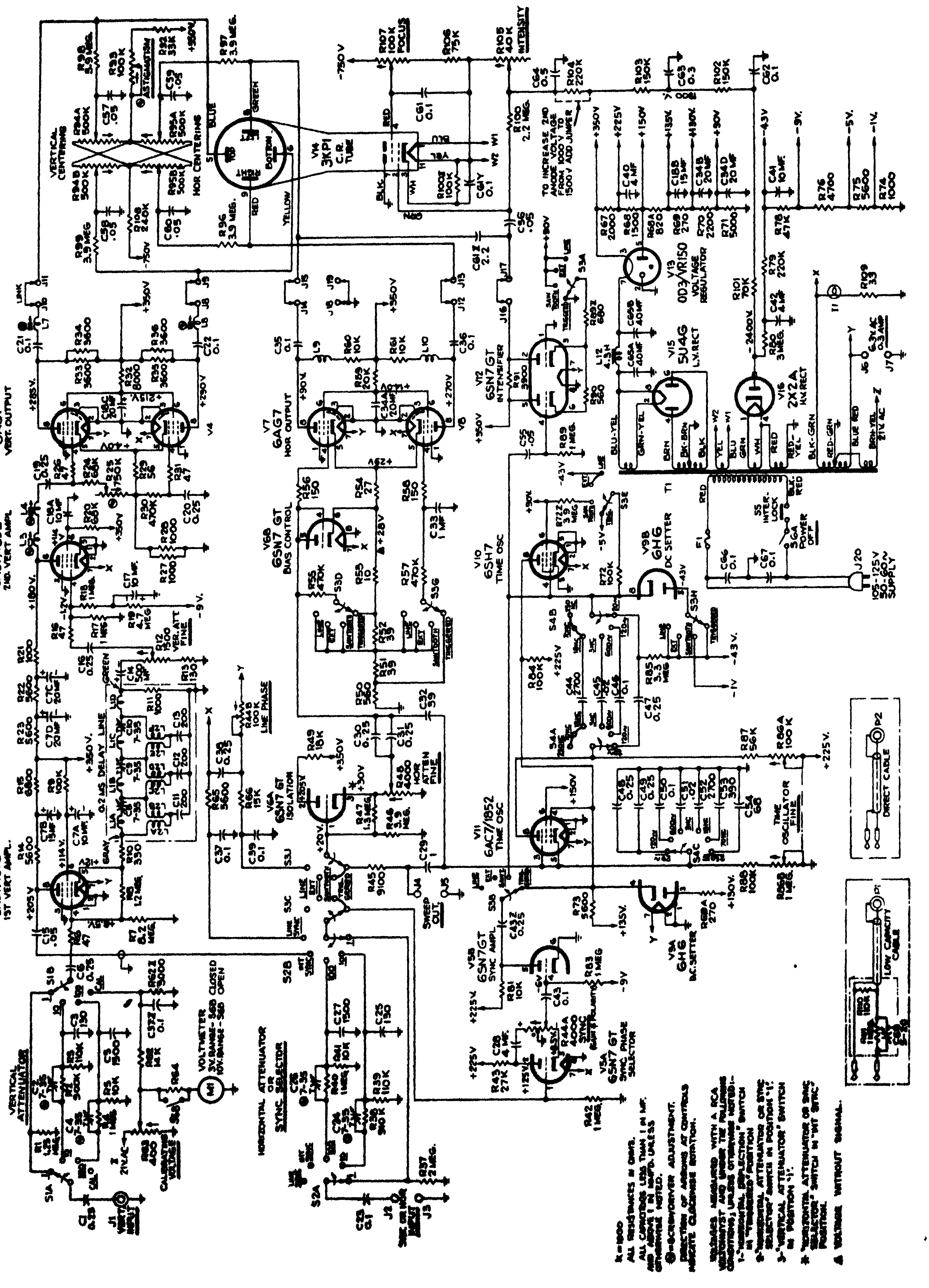
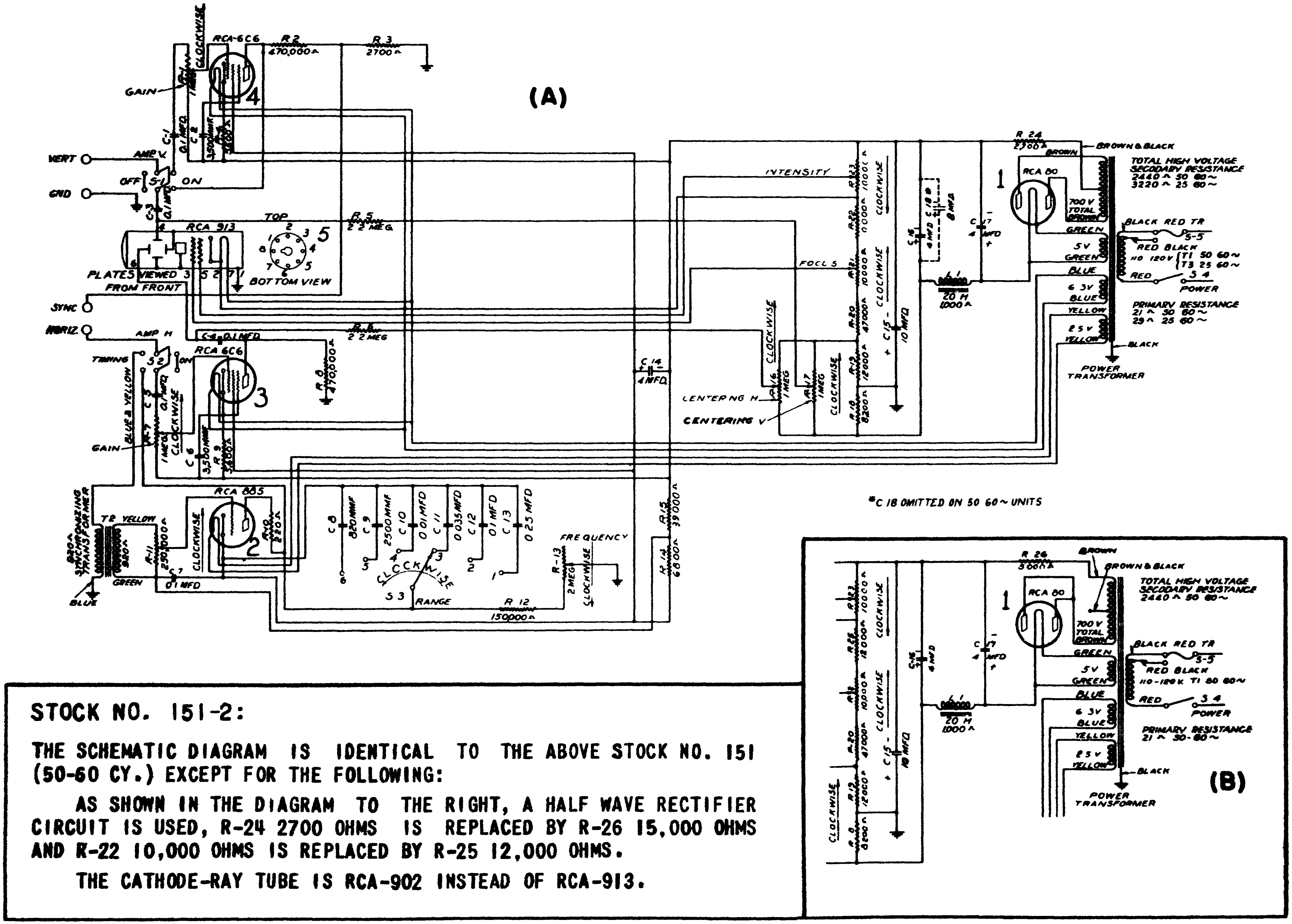

V1 6AC7/1852 1ST VERT AMPL.
V2 6AC7/1852 2ND VERT AMPL.
V3 6AG7 VERT OUTPUT
V4
V5A 6SN7 GT SYNC PHASE SELECTOR
V5B 6SN7/GT SYNC AMPL.
V6A 6SN7 GT ISOLATION
V6B 6SN7 GT BIAS CONTROL
V7 6AG7 HOR OUTPUT
V8
V9A 6H6 D.C. SETTER
V9B 6H6 DC SETTER
V10 6SJ7 TIME OSC
V11 6AC7/1852 TIME OSC
V12 6SN7/GT INTENSIFIER
V13 OD3/VR150 VOLTAGE REGULATOR
V14 3KP1 C.R. TUBE
V15 5U4G L.V. RECT
V16 2X2A H.V. RECT
VERTICAL ATTENUATOR
CALIBRATING VOLTAGE
VOLTMETER
HORIZONTAL ATTENUATOR OR SYNC SELECTOR
0.2 µS DELAY LINE
VERTICAL CENTERING
HOR CENTERING
ASTIGMATISM
FOCUS
INTENSITY
TO INCREASE 2ND ANODE VOLTAGE FROM 1000 V TO 1500 V ADD JUMPER
LINE PHASE
HOR ATTEN FINE
VER. ATT FINE
TIME OSCILLATOR FINE
SWEEP OUT
105-125 V 50-60~ SUPPLY
POWER OFF
INTER-LOCK
DIRECT CABLE
LOW CAPACITY CABLE
K=1000
ALL RESISTANCES IN OHMS.
ALL CAPACITORS LESS THAN 1 IN MF. AND ABOVE 1 IN MMFD. UNLESS OTHERWISE NOTED.
⊕=SCREWDRIVER ADJUSTMENT.
DIRECTION OF ARROWS AT CONTROLS INDICATE CLOCKWISE ROTATION.
VOLTAGES MEASURED WITH A RCA VOLTOHMYST AND UNDER THE FOLLOWING CONDITIONS, UNLESS OTHERWISE NOTED:-
1-"HORIZONTAL DEFLECTION" SWITCH IN "TRIGGERED" POSITION
2-"HORIZONTAL ATTENUATOR OR SYNC SELECTOR" SWITCH IN POSITION "1".
3-"VERTICAL ATTENUATOR" SWITCH IN POSITION "1".
* "HORIZONTAL ATTENUATOR OR SYNC SELECTOR" SWITCH IN "INT SYNC" POSITION.
▲ VOLTAGE WITHOUT SIGNAL.

Fig. 22-51.—Schematic of RCA Model WO-79A.

Courtesy RCA



STOCK NO. 151-2:

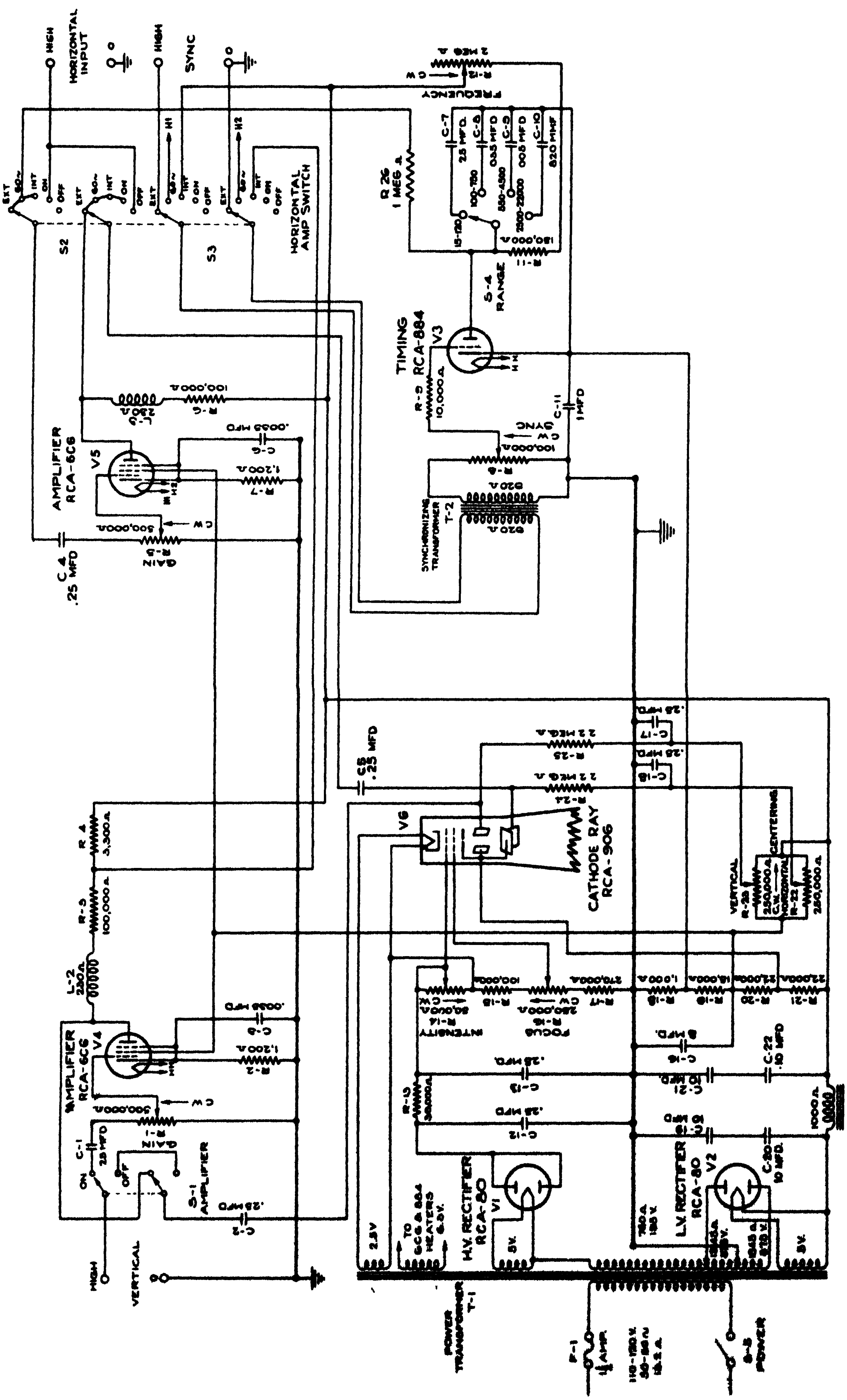
THE SCHEMATIC DIAGRAM IS IDENTICAL TO THE ABOVE STOCK NO. 151 (50-60 CY.) EXCEPT FOR THE FOLLOWING:

AS SHOWN IN THE DIAGRAM TO THE RIGHT, A HALF WAVE RECTIFIER CIRCUIT IS USED, R-24 2700 OHMS IS REPLACED BY R-26 15,000 OHMS AND R-22 10,000 OHMS IS REPLACED BY R-25 12,000 OHMS.

THE CATHODE-RAY TUBE IS RCA-902 INSTEAD OF RCA-913.

Fig. 22-52.—(A) Schematic of RCA Model 151, 151-A, (B) Model 151-2.

Courtesy RCA



Courtesy RCA

Fig. 22-53.—Schematic of RCA Model 155-A.

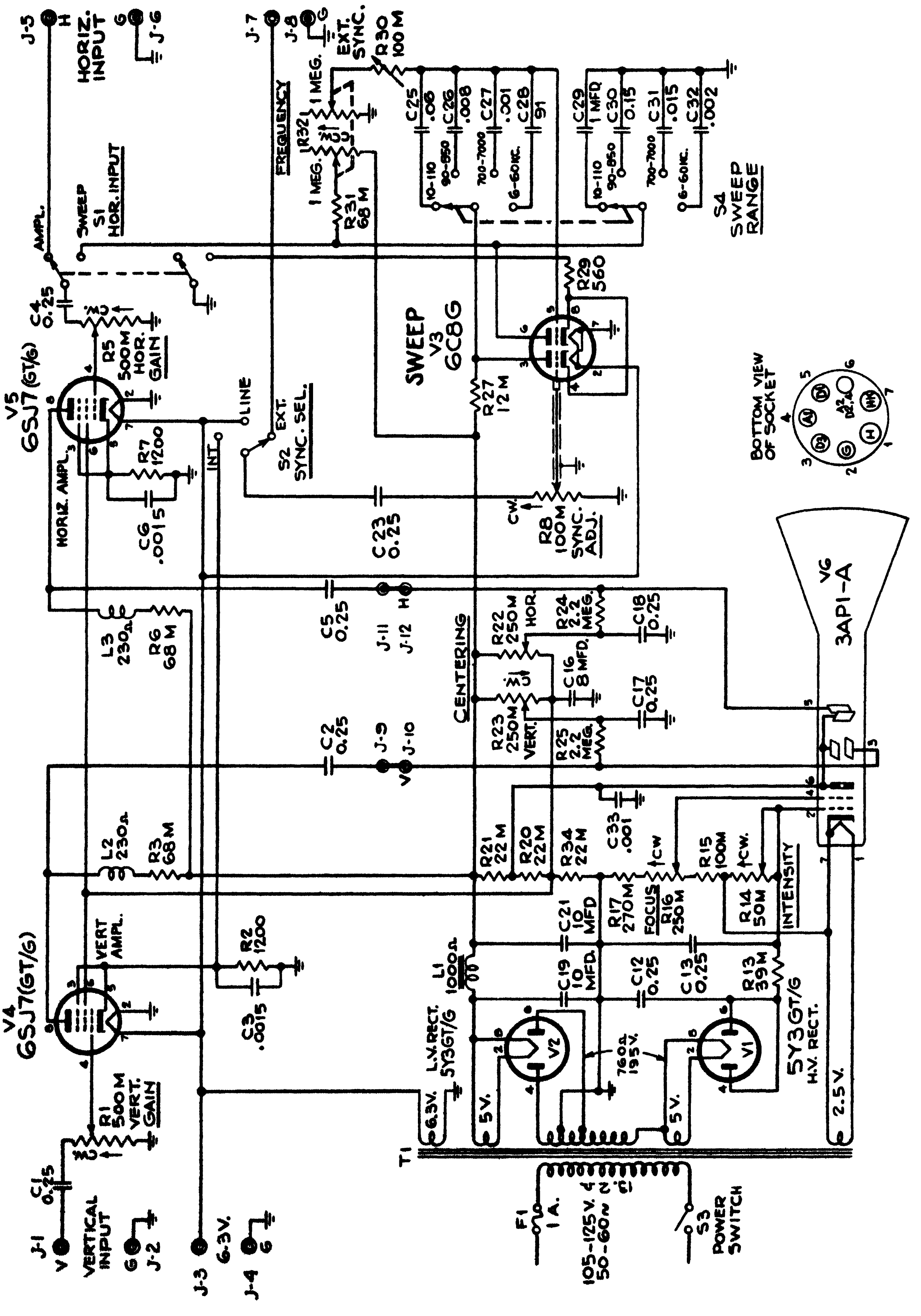


Fig. 22-54.—Schematic of RCA Model 155-C.

Courtesy RCA

RCA MODEL 158

Frequency Response
Vertical Amplifier 15 cps to 200 kc, ± 10%
Horizontal Amplifier 10 cps to 30 kc, ± 10%
Sweep Circuit 4 cps to 18 kc

Deflection Factors
Vertical Amplifier 0.04 rms volts/inch
Vertical Amplifier through Probe 0.4 rms volts/inch
Horizontal Amplifier 0.75 rms volts/inch
Horizontal-Deflection Plates 17.5 rms volts/inch

Line Rating 105-125 volts, 50-60 cps

Tube Complement

Type	*Function*
6C6 (V1)	Horizontal Amplifier
6SJ7 (V2)	Vertical Amplifier, First Stage
6AC7 (V3)	Vertical Amplifier, Second Stage
884 (V4)	Sawtooth-Sweep Oscillator
5BP1 or 5HPI (V5)	Cathode-Ray Tube
80 (V6)	Low-Voltage Rectifier
879 (V7)	High-Voltage Rectifier

The schematic circuit diagram for Model 158 is shown in Fig. 22-55. A special high-resistance frequency-compensated four-step attenuator is employed between the vertical-input terminals and the vertical amplifiers. A 10,000-ohm preset potentiometer *R-29* which is in the coupling circuit between the 6SJ7 and 6AC7 vertical amplifiers is used as a low-frequency adjustment.

RCA MODEL 160-B

Frequency Response
Vertical Amplifier 15 cps to 12,000 cps
Horizontal Amplifier 10 cps to 30,000 cps
Sweep Circuit 4 cps to 18,000 cps

Deflection Factors
Vertical Amplifier 0.02 rms volts/inch
Vertical-Deflection Plates 17.5 rms volts/inch
Horizontal Amplifier 0.75 rms volts/inch
Horizontal-Deflection Plates 17.5 rms volts/inch

Line Rating 105-125 volts, 50-60 cps

Tube Complement

Type	*Function*
6C6 (VI)	Horizontal Amplifier
6C6 (V2, V3)	Vertical Amplifiers
884 (V4)	Sweep Oscillator
5BP1/1802-P1 (V5) or 5HP1	Cathode-Ray Tube
80 (V6)	Low-Voltage Rectifier
879 (V7)	High-Voltage Rectifier
VR-105-30 (V8)	Voltage Regulator
VR-150-30 (V9)	Voltage Regulator

The schematic circuit diagram of Model 160-B is shown in Fig. 22-56. The circuit design is not particularly complex. A somewhat unusual feature is the use of two gaseous voltage regulator tubes, connected in series, to stabilize the screen potential of *V2*. A vertical-deflection reversing switch *S6* is connected to the vertical-deflection plates so that the polarity of vertical deflection may be changed at will. This is useful when observing resonance curves where the detector polarity may invert the curve.

RCA MODEL 304-A

Frequency Response
Vertical Amplifier 4 cps to 100 kc
Horizontal Amplifier 4 cps to 100 kc
Sweep Circuit 4 cps to 18 kc

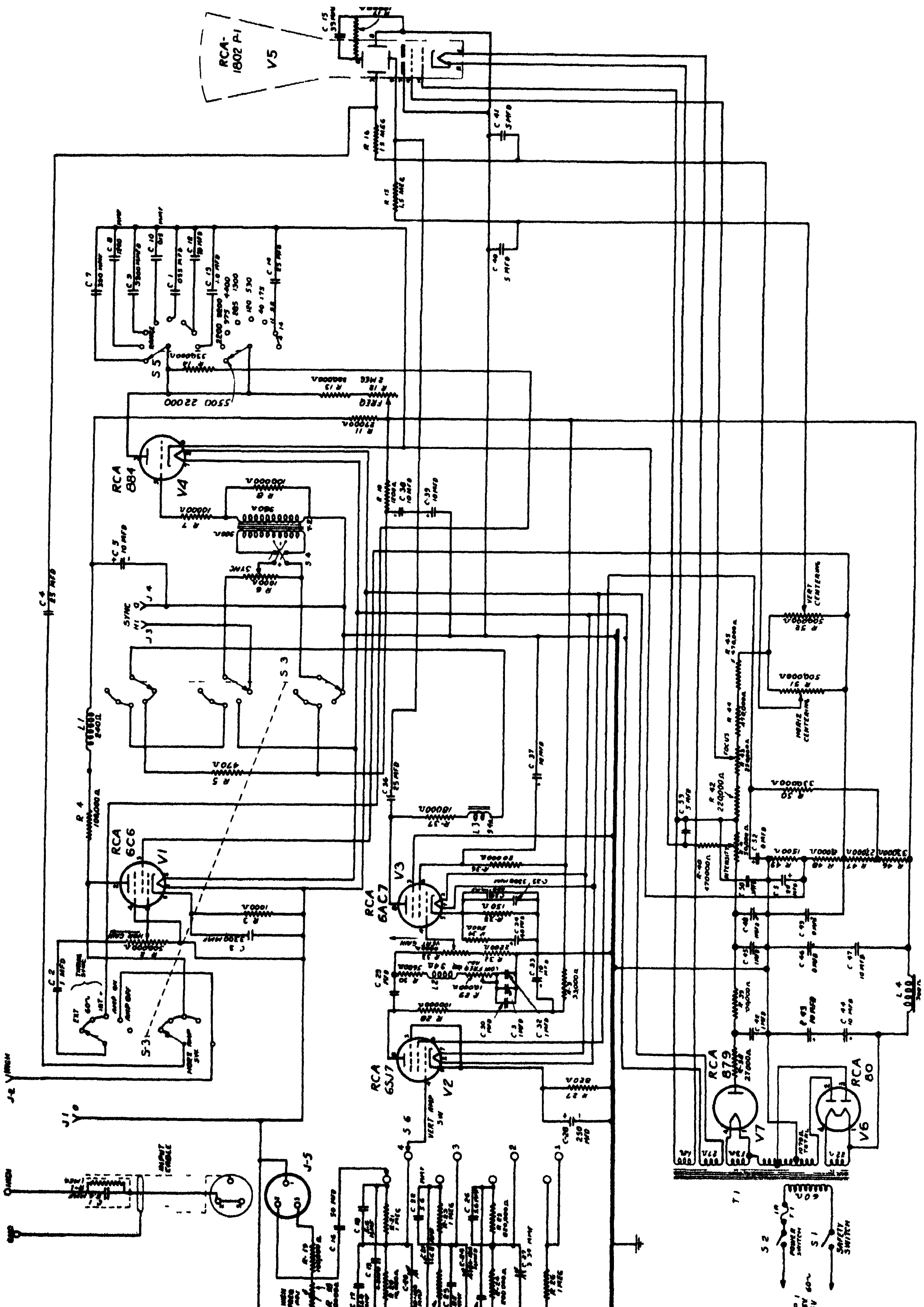
Deflection Factors
Vertical Amplifier 0.02 rms volts/inch
Vertical-Deflection Plates 29 rms volts/inch
Horizontal Amplifier 0.02 rms volts/inch
Horizontal-Deflection Plates 29 rms volts/inch

Line Rating 110-120 volts, 50-60 cps

Tube Complement

Type	*Function*
6C (1, 2)	Vertical Amplifier
6C6 (3, 4)	Vertical-Output Deflection
6C6 (5, 6)	Horizontal Amplifier
6C6 (7, 8)	Horizontal-Output Deflection
6J7 (9)	Sync Amplifier
885 (10)	Gaseous-Sweep Oscillator
6N7G (11)	Blanking Amplifier
879 (12)	High-Voltage Rectifier
2A3 (13)	Voltage Regulator
6J7 (14)	Control Tube
874 (15)	Voltage Regulator
5Z3 (16)	Low-Voltage Rectifier
914 (17)	Cathode-Ray Tube

The schematic circuit diagram for Model 304-A which employs a 9-inch cathode-ray tube, is shown in Fig. 22-57. Both deflection amplifiers are identical and each will accommodate either push-pull (three-wire, balanced to ground) or single-ended (two-wire, unbalanced) input circuits. A blanking circuit is employed, tube *11*, which operates as a two-stage clipping amplifier, the output of which is capacitance-coupled to the control grid of the cathode-ray tube. The output of this amplifier is used to change the bias on the cathode-ray tube, just as the intensity control does, and thus modulate the intensity of the beam, for time or frequency determination.



Courtesy RCA

Fig. 22-55.—Schematic of RCA Model 158.

Courtesy RCA

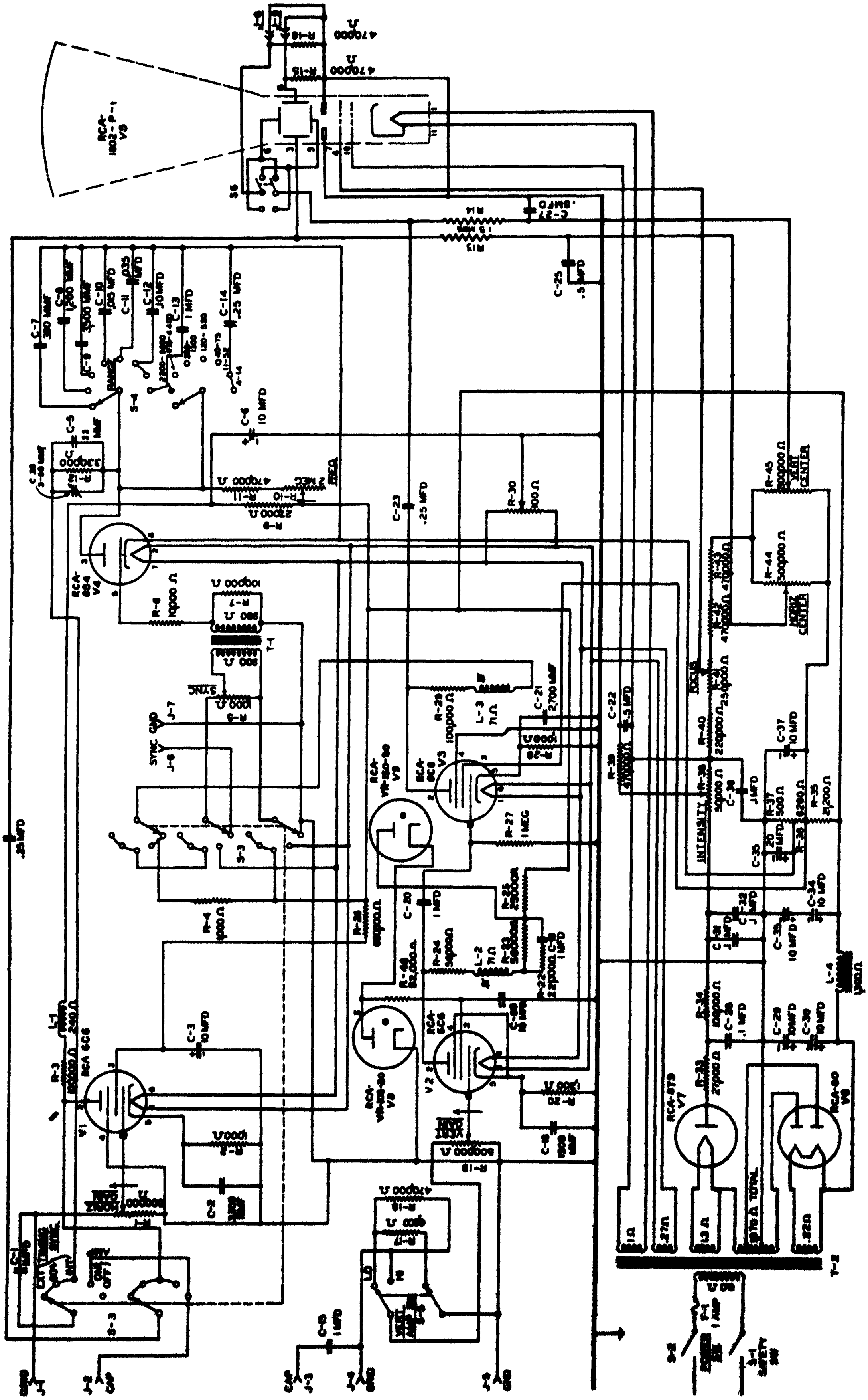
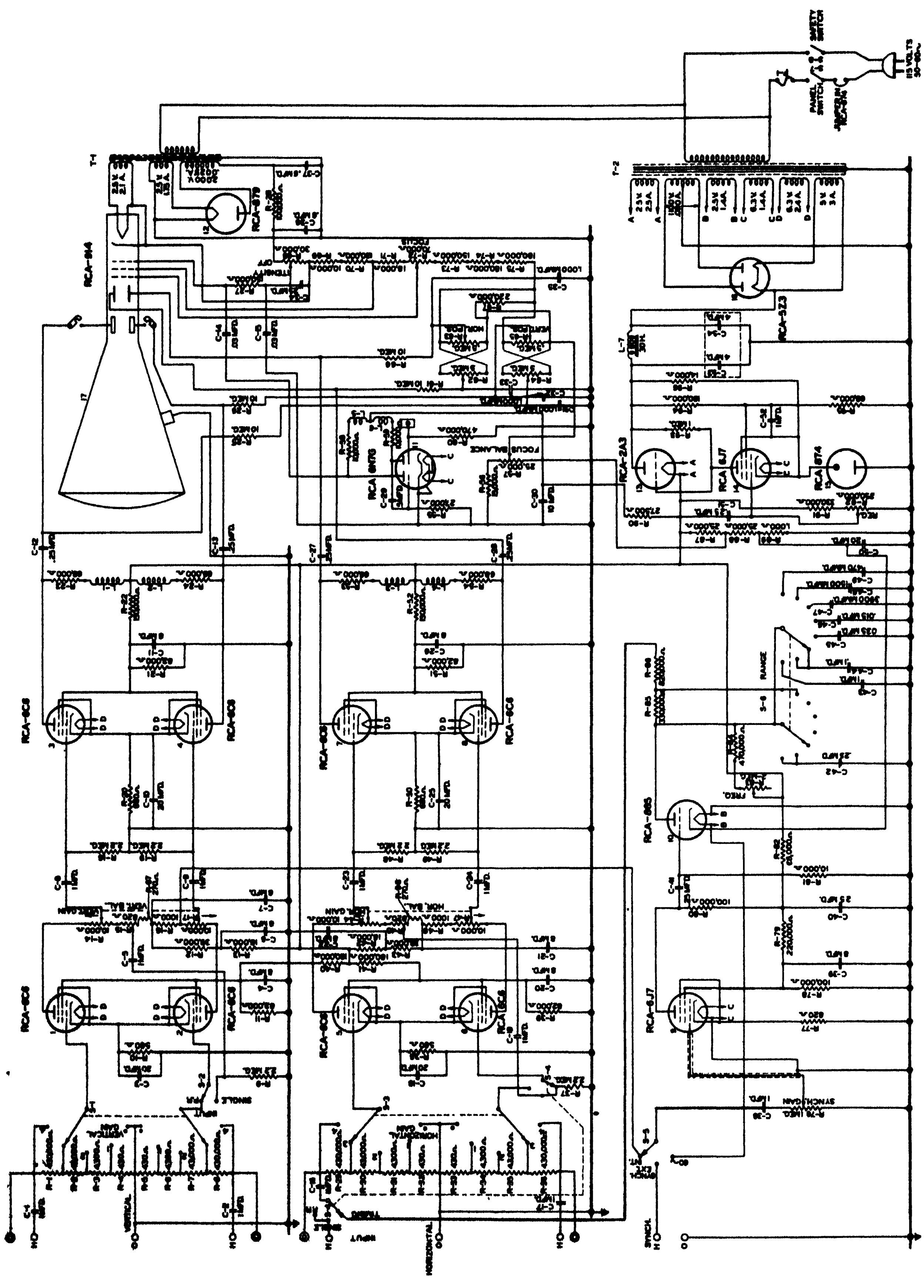
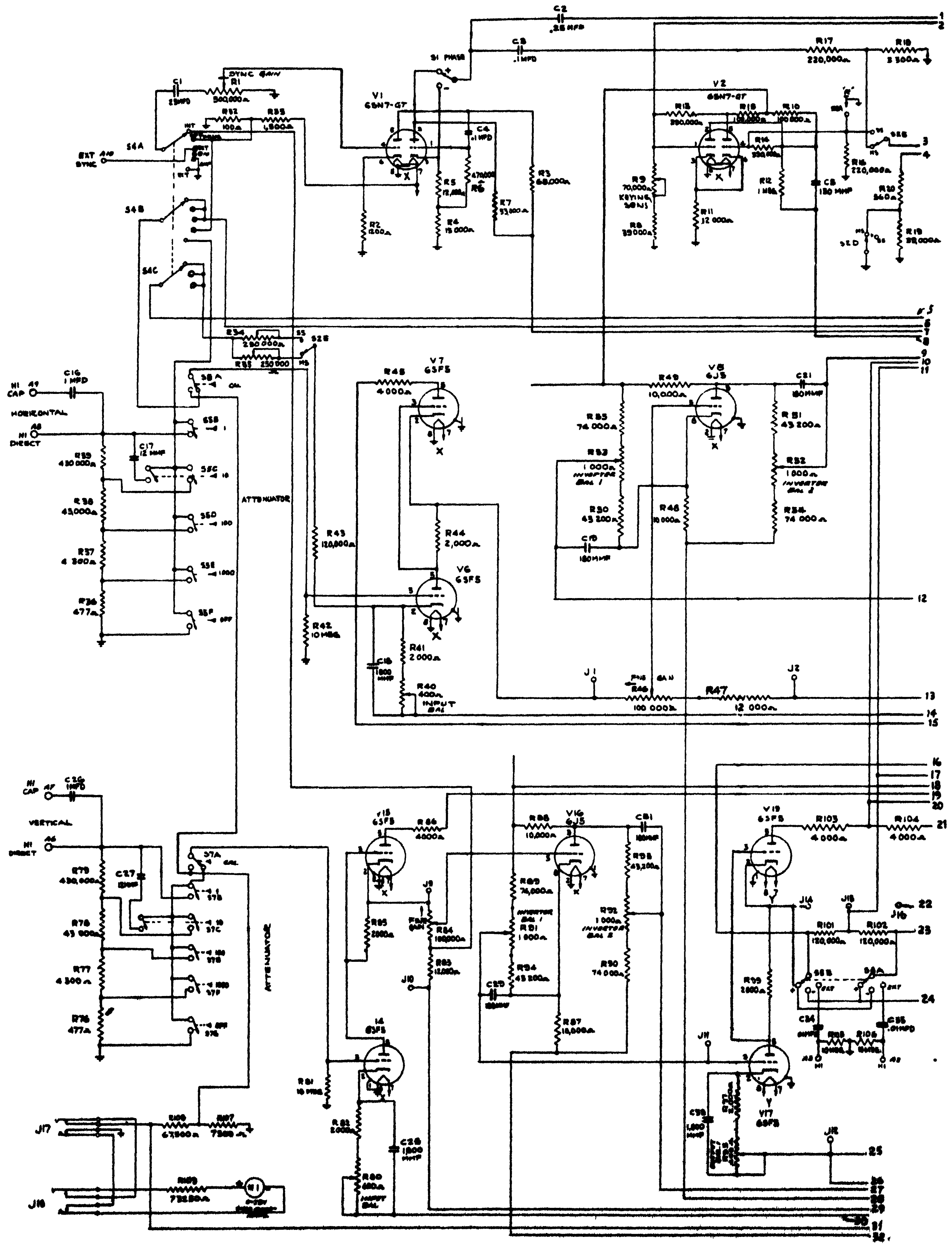


Fig. 22-56.—Schematic of RCA Model 160-B.



Courtesy RCA

Fig. 22-57.—Schematic of RCA Model 304-A.


C2
.25 MFD
C3
.1 MFD
S1 PHASE
R17
220,000Ω
R18
3300Ω
C1
25 MFD
SYNC GAIN
R1
500,000Ω
R52
100Ω
R53
1,500Ω
S4A
EXT SYNC
S4B
S4C
V1
6SN7-GT
C4
.1 MFD
R5
R6
R7
R3
68,000Ω
R2
1200Ω
R4
15 000Ω
V2
6SN7-GT
R13
R15
R10
R14
R9
70,000Ω
KEYING SENS
R8
39000Ω
R11
12 000Ω
R12
1 MEG
C5
150 MMF
R16
220,000Ω
R20
560Ω
R19
39,000Ω
SED
R34
250 000Ω
R35
250 000
S5
S2E
V7
6SF5
R45
4000Ω
V8
6J5
R49
10,000Ω
C21
150MMF
R55
74 000Ω
R51
43 200Ω
R53
1 000Ω
INVERTER BAL 1
R52
1 000Ω
INVERTER BAL 2
R50
43 200Ω
R48
10 000Ω
R54
74 000Ω
C19
150MMF
HI CAP
C16
1 MFD
HORIZONTAL
HI DIRECT
C17
12 MMF
S5A
CAL
S5B
S5C
S5D
S5E
S5F
OFF
R39
430 000Ω
R38
43,000Ω
R37
4 300Ω
R36
477Ω
ATTENUATOR
R43
120,000Ω
R44
2,000Ω
V6
6SF5
R42
10 MEG
C18
1800 MMF
R41
2 000Ω
R40
400Ω
INPUT BAL
J1
J2
R46
GAIN
100 000Ω
R47
12 000Ω
HI CAP
C26
1 MFD
VERTICAL
HI DIRECT
R79
430,000Ω
C27
12MMF
S7A
CAL
S7B
S7C
S7D
S7E
S7F
OFF
R78
43 000Ω
R77
4 300Ω
R76
477Ω
ATTENUATOR
V15
6SF5
R86
4000Ω
J9
FINE GAIN
R84
100,000Ω
R85
2000Ω
R83
12,000Ω
J10
V16
6J5
R88
10,000Ω
C31
150MMF
R89
74,000Ω
INVERTER BAL 1
R91
1 000Ω
R94
43 200Ω
C29
150MMF
R93
43,200Ω
R92
1 000Ω
INVERTER BAL 2
R90
74 000Ω
R87
10,000Ω
V19
6SF5
R103
4 000Ω
R104
4 000Ω
J14
J15
J16
R101
120,000Ω
R102
120,000Ω
S6B
S6A
EXT
R99
2000Ω
C34
C35
.01MFD
R105
R106
10 MEG
J11
V17
6SF5
C33
1,800 MMF
J12
V14
6SF5
R81
10 MEG
R82
2000Ω
C28
1800 MMF
R80
400Ω
INPUT BAL
J17
R100
67,500Ω
R107
7500Ω
J18
R108
75,250Ω
M1
1
2
3
4
5
6
7
8
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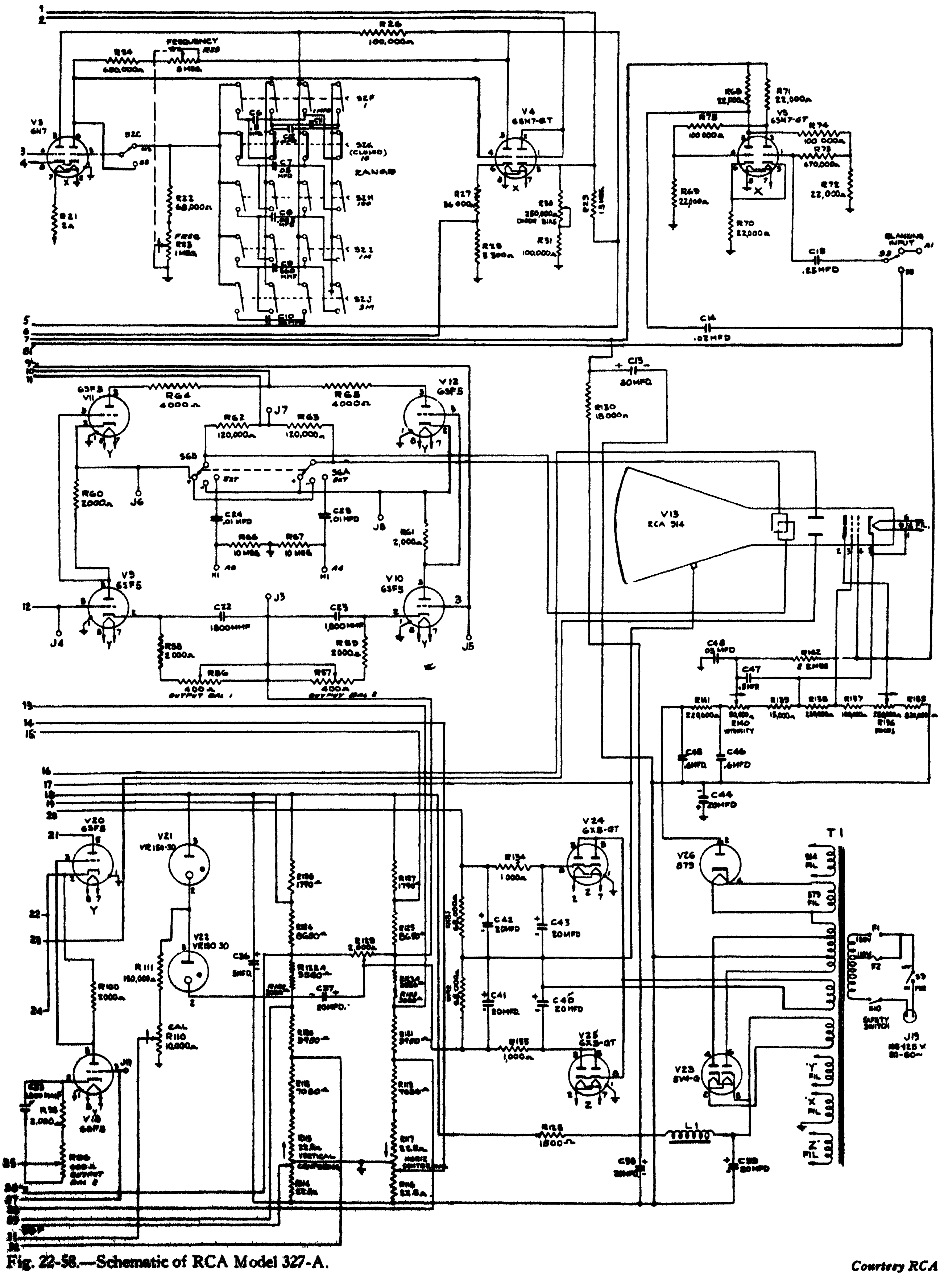


Fig. 22-58.—Schematic of RCA Model 327-A.

Courtesy RCA

RCA MODEL 327-A

FREQUENCY RESPONSE
Vertical Amplifier 0 cps to 100 kc
Horizontal Amplifier 0 cps to 100 kc
Sweep Circuit 1 cps to 30 kc
Blanking Amplifier 30 cps to 100 kc

DEFLECTION SENSITIVITY
Vertical Amplifier 0.025 rms volts/inch
Vertical-Deflection Plates 52 rms volts/inch
Horizontal Amplifier 0.03 rms volts/inch
Horizontal-Deflection Plates 52 rms volts/inch

LINE RATING 105-115/115-125/ and 210-230/230-250 volts, 50/60 cps

TUBE COMPLEMENT

Type	*Function*
914	Cathode-Ray Tube
6SF5 (twelve)	Voltage Amplifiers
6J5 (two)	Voltage Amplifier (Phase Inverter)
6SN7GT (four)	Sync Amplifier, Keying, Rectifier, Blanking Amplifier
6N7	Timing-Axis (Sweep) Oscillator
879	High-Voltage Rectifier
5V4G	Low-Voltage Rectifier
6X5GT (two)	Voltage-Doubler Rectifiers
VR150-30 (two)	Voltage Regulators

The schematic circuit diagram for Model 327-A is shown in Fig. 22-58.

Timing Oscillator

The timing oscillator circuit uses a special circuit which functions as a one shot multivibrator, or multiple-sweep operation. A 6N7 *(V3)* is used in the timing oscillator circuit. The 6N7 is also a part of the single-sweep circuit. The output of the timing-axis oscillator is fed to a 6SN7 cathode-follower tube *V4*, whose output in turn is delivered to the 6SF5 *(V6)* horizontal amplifier. The single-sweep rate corresponds in frequency to that obtained in the normal multi-sweep operation. The single-sweep circuit uses three double-triode tubes: the timing axis tube *V3*, a double triode as a keying tube *V2*, and the cathode-follower-triode section of a third tube *V4* that has its other section connected as a biased diode. The biased-diode section of *V4* is the first tube in the circuit and permits signals of but one polarity to be applied to the keying tube *V2*.

The action of the single-sweep circuit is as follows: When a signal is applied to the vertical amplifier, a voltage appears at pin 4 of the 6SN7GT synchronizing amplifier *V1*. This voltage is taken from the output of *V15*, at the junction of *R83*. The voltage appearing at the plate of *V1* is coupled through capacitor *C4* to pin 1 of the second-triode section of this tube. This section of the tube is used as phase inverter for the purpose of reversing the phase of the signal applied to the biased diode section of *V4* which has pins 1-4, grid and plate, tied together for single-sweep operation, and for normal synchronizing phase when using multisweep.

In the starting condition for sweep, the first-triode section of the keying tube *V2* (pins 1, 2, 3) is conducting, and the second-triode section is cut off. The rectified current passed by the diode section of *V4* is of such polarity as to apply a negative voltage to pin 1 of *V2*. The application of this negative voltage causes the first-triode section to become nonconducting. With the second triode conducting, a high positive d-c voltage appears across *R16*, the grid resistor for the first-triode section of the sweep oscillator *V3*.

Since both triode sections of *V3* have a common cathode resistance, the operation potential of the diode plate (formed by connecting the plate and grid together by switch *S2C*) is controlled by the drop in this resistance *(R19* and *R20)*. Changes of bias on the input grid of *V3*, pin 4, then causes the diode section to be conducting when the cathode potential rises, and nonconducting when it falls. Since a capacitator is connected in the diode circuit, the rate of change in voltage is controlled by the resistance and capacitance in the circuit *(R22, R23* and any of the range capacitors, *C6, C7, C8,* or *C10*, depending on which range button is pressed). In multiple sweep, these same components are used, the single-sweep time corresponding to a single across-the-screen movement of the multiple sweep. This sweep output is impressed on pin 4 of the cathode-follower tube *V4*, where the output taken from its cathode circuit is impressed on the input of the horizontal amplifier (grid of *V6*). Movement of the spot on the screen of the cathode-ray tube then results. To prevent a spot from appearing before the start of the sweep, the blanking amplifier *V5* is connected to the keying tube *V2*.

Vertical Output Amplifier

A study of the circuit of the output stage, *V17, V18, V19,* and *V20*, shows that this is a push-pull type. It is, however, distinctly unusual because d-c potentials are applied to the grids and because series operation is used for each half of the stage. This can be seen by tracing from *R95* to *R97, V17, R99, V19, R103* and *B+*, so that it is apparent that *V17-V19* form a series circuit; and from *R96* to *R98, V18, R100, V20, R104*, and *B+*. *V18* and *V20* also form a series circuit. The two series circuits, considered together form a push-pull amplifier stage.

The power-output stage or deflection-stage circuit is quite complicated and cannot be discussed in detail here. It is really a differential d-c push-pull amplifier, consisting basically of a bridge circuit formed by arms R101, R102, and R103, R104. The amount of unbalanced current flow, which is coupled to the deflection plates, depends on the relative input signal.

RCA MODEL 715-B

FREQUENCY RESPONSE
Vertical Amplifier 5 cps to 11 Mc, flat within ± 1 db
Horizontal Amplifier 3 cps to 500 kc, flat within ± 2 db
Sweep Circuit Sawtooth Range, 5 cps to 100 kc, Sync Range, 5 cps to 2 Mc

DEFLECTION FACTORS
Vertical Amplifier 0.1 rms volts/inch, (2nd anode 2,600 V)
Vertical-Deflection Plates 35 rms volts/inch, (2nd anode, 2,600 V)
Horizontal Amplifier 0.5 rms volts/inch, (2nd anode 2,600 V)
Horizontal-Deflection Plates 29 rms volts/inch, (2nd anode 2,600 V)

LINE RATING 105-125/210-250 volts, 50-60 cps

The schematic circuit diagram including tube complement and functions is shown in Figs. 22-59A (Power Supply) and 22-59B (Oscilloscope). This oscilloscope possesses a number of unusual features, which will be discussed.

Vertical Amplifier

The input to the vertical amplifier is fed through a five-step attenuator unit. The input amplifier uses high-frequency compensation in form of *L101* which is in series with its plate circuit. The 6V6 push-pull vertical-driver stage *V105* is unusual because of the frequency compensation network arrangement. Series inductance *L105* tends to attenuate high-frequency signals, serving as a low-pass filter, while *L107A* in series with the plate-load resistor *R150*, tends to boost the voltage output at the higher frequencies. A similar action occurs with respect to the other half of the 6V6 vertical-driver stage *V106*. The net effect involves differential action between the various components, the shape of the response curve for this stage being alterable by means of the variable inductances *L105, L107A, L107B,* and *L106*. The over-all curve for the entire vertical amplifier is a summation of the response characteristics of the individual stages. The push-pull 6V6 driven stage excites the push-pull 807 vertical-output stage, *V107* and *V108*. The 807 plate circuits are also frequency compensated to give a flat over-all, equalized response curve.

Horizontal Amplifier

The output stage provides low-frequency boost by means of *L119* and *L120* in the plate circuit of *V114* and *V115*, respectively. In addition to this low-frequency compensation, there is a high-frequency boost provided by *L117* and *L118* to compensate for high-frequency attenuation. The spot positioning control circuit is driven by the push-pull 6AG7 horizontal-output.

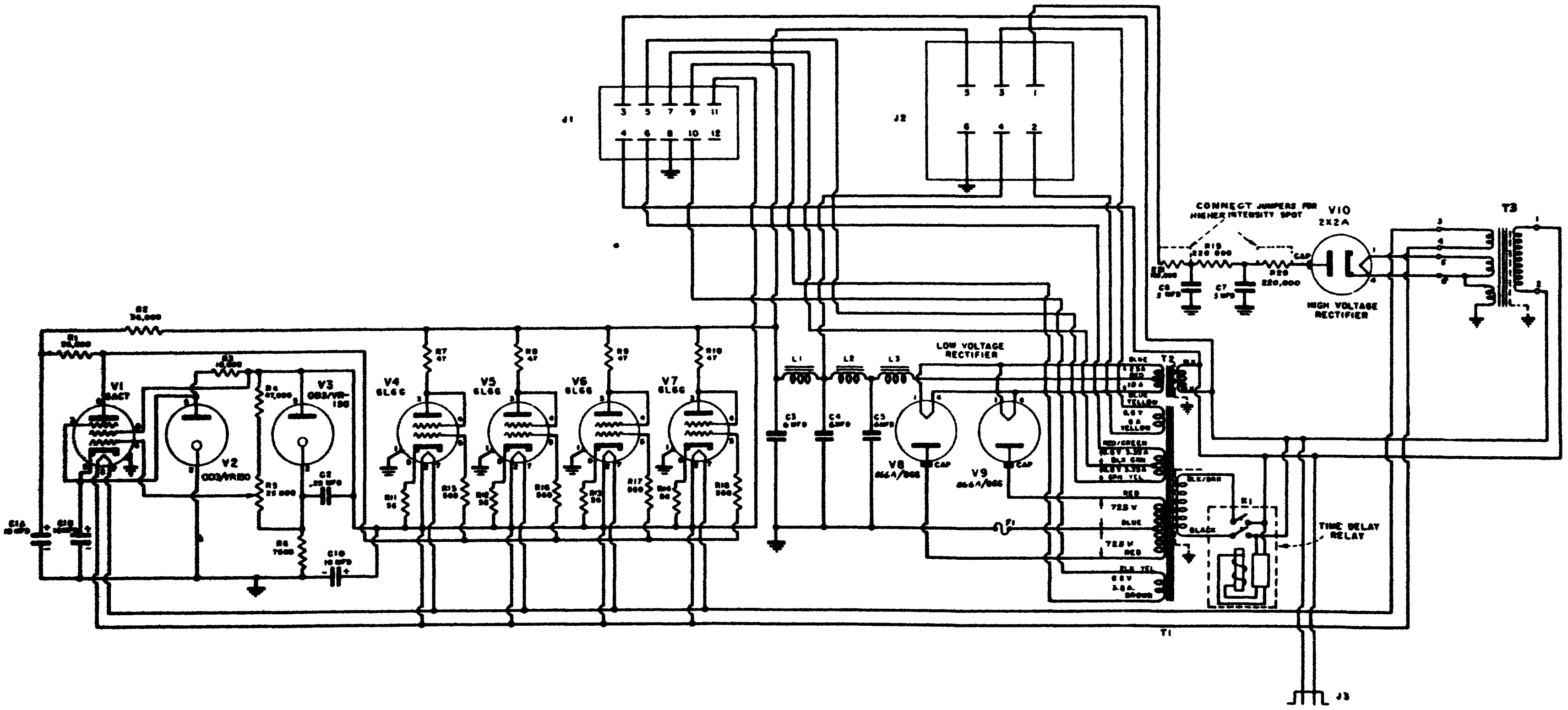


Fig. 22-59A.—Power supply of RCA Model 715-B.

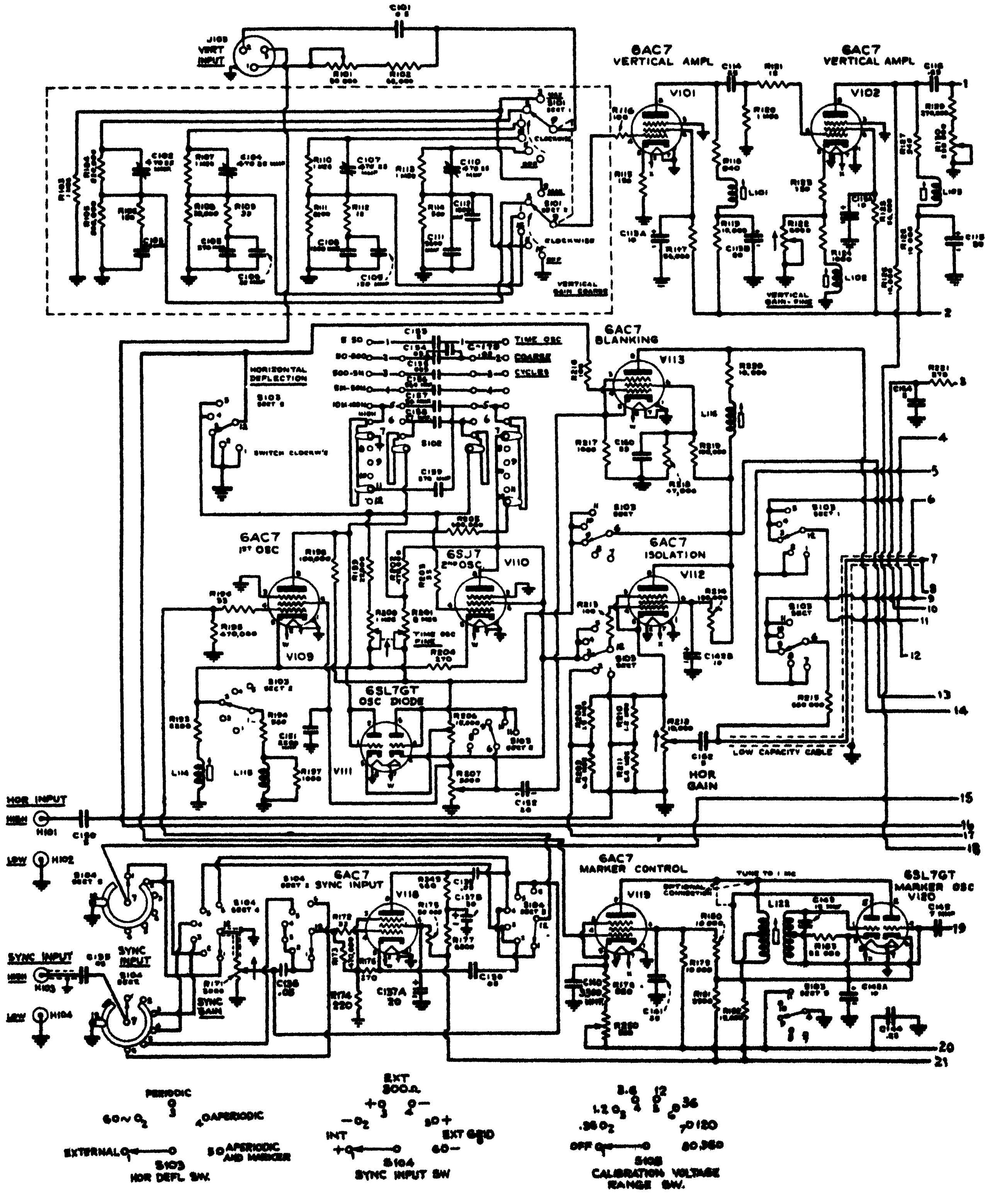
Courtesy RCA

Marker Control

The marker control tube *V119* receives its excitation from the grid circuit of the blanking amplifier *V113*. When *S103*, the horizontal-deflection switch is in the fifth, or APERIODIC AND MARKER, position, timing marker pulses are superimposed on the signal voltage. This produces a visible indication on the signal trace at selected time intervals. In the type 715-B oscilloscopes, the intervals have been set at one microsecond. The use of the marker-control tube, which is keyed by the sweep-oscillator voltage, insures that the timing markers will appear stationary on the screen of the cathode-ray tube.

Oscillator Diode

In the fourth, or APERIODIC, position of the HOR DEFLECTION control, the Potter Oscillator circuit is converted from a repetitive oscillator to an aperiodic or "one-shot" type for examination of waveforms that do not recur at regular intervals. The change is accomplished by increasing the grid bias on both of the oscillator tubes. When the oscillator is triggered by a sync pulse, a single cycle of oscillation occurs, causing the spot to sweep across the screen of the cathode-ray tube and return to its starting position. It remains there until the oscillator is triggered by the next sync pulse.


J103 VERT INPUT
C101
R101
R102
6AC7 VERTICAL AMPL
V101
6AC7 VERTICAL AMPL
V102
VERTICAL GAIN COARSE
VERTICAL GAIN-FINE
CLOCKWISE
L101
L102
L103
6AC7 BLANKING
V113
TIME OSC COARSE CYCLES
HORIZONTAL DEFLECTION
S103
SWITCH CLOCKW'S
S102
6AC7 1ST OSC
V109
6SJ7 2ND OSC
V110
TIME OSC FINE
6AC7 ISOLATION
V112
6SL7GT OSC DIODE
V111
HOR GAIN
LOW CAPACITY CABLE
HOR INPUT
HIGH
H101
LOW
H102
S104
SYNC INPUT
H103
H104
SYNC GAIN
6AC7 SYNC INPUT
V116
6AC7 MARKER CONTROL
V119
OPTIONAL CONNECTION
TUNE TO 1 MC
L121
6SL7GT MARKER OSC
V120
1
2
3
4
5
6
7
8
9
10
11
12
13
14
15
16
17
18
19
20
21
PERIODIC
60~
APERIODIC
EXTERNAL
APERIODIC AND MARKER
S103 HOR DEFL SW.
EXT 500Ω
INT
EXT GRID
S104 SYNC INPUT SW
3.6
12
1.2
36
.36
120
OFF
360
S105 CALIBRATION VOLTAGE RANGE SW.

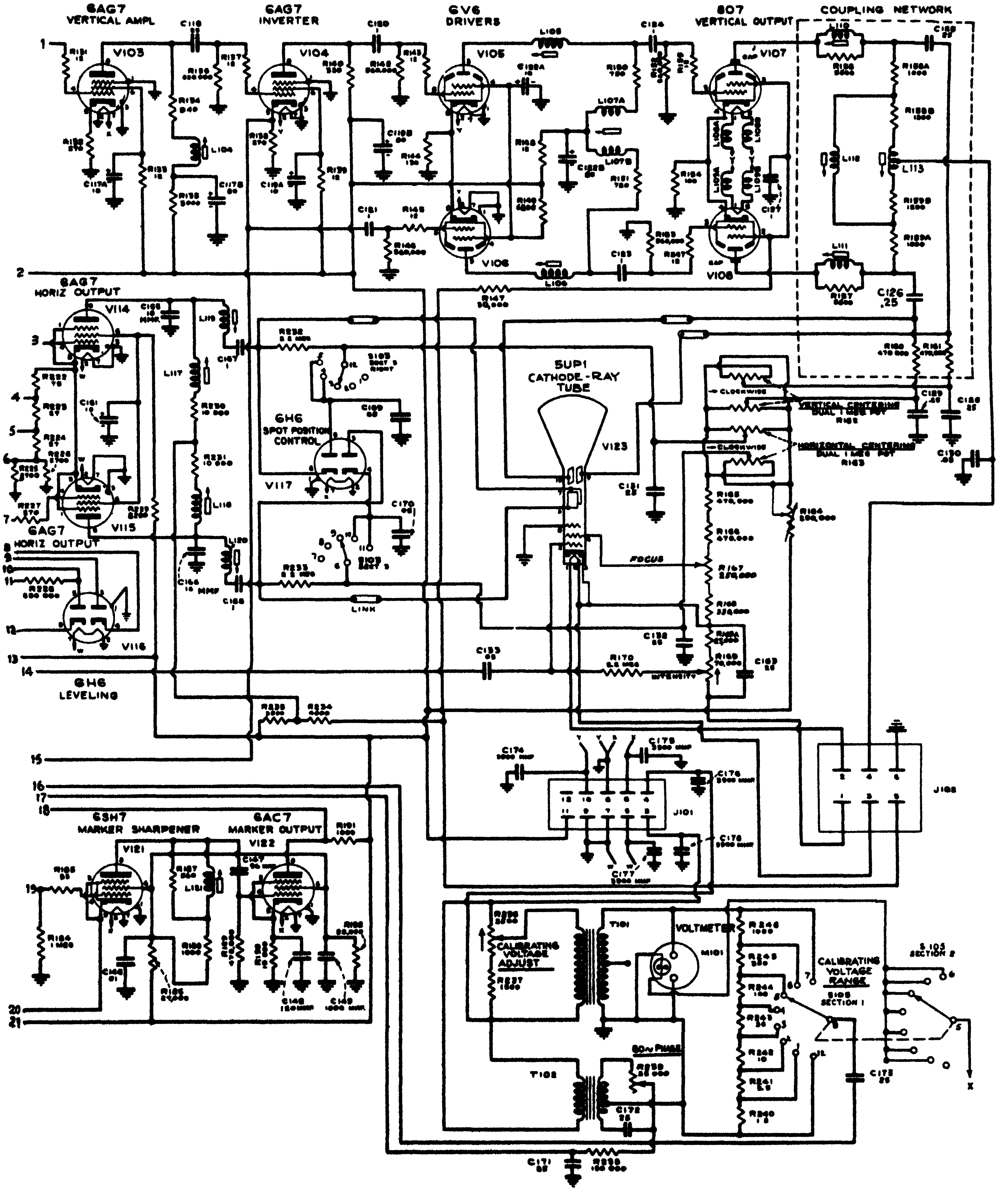


Fig. 22-59B.—Schematic of RCA Model 715-B.

Courtesy RCA

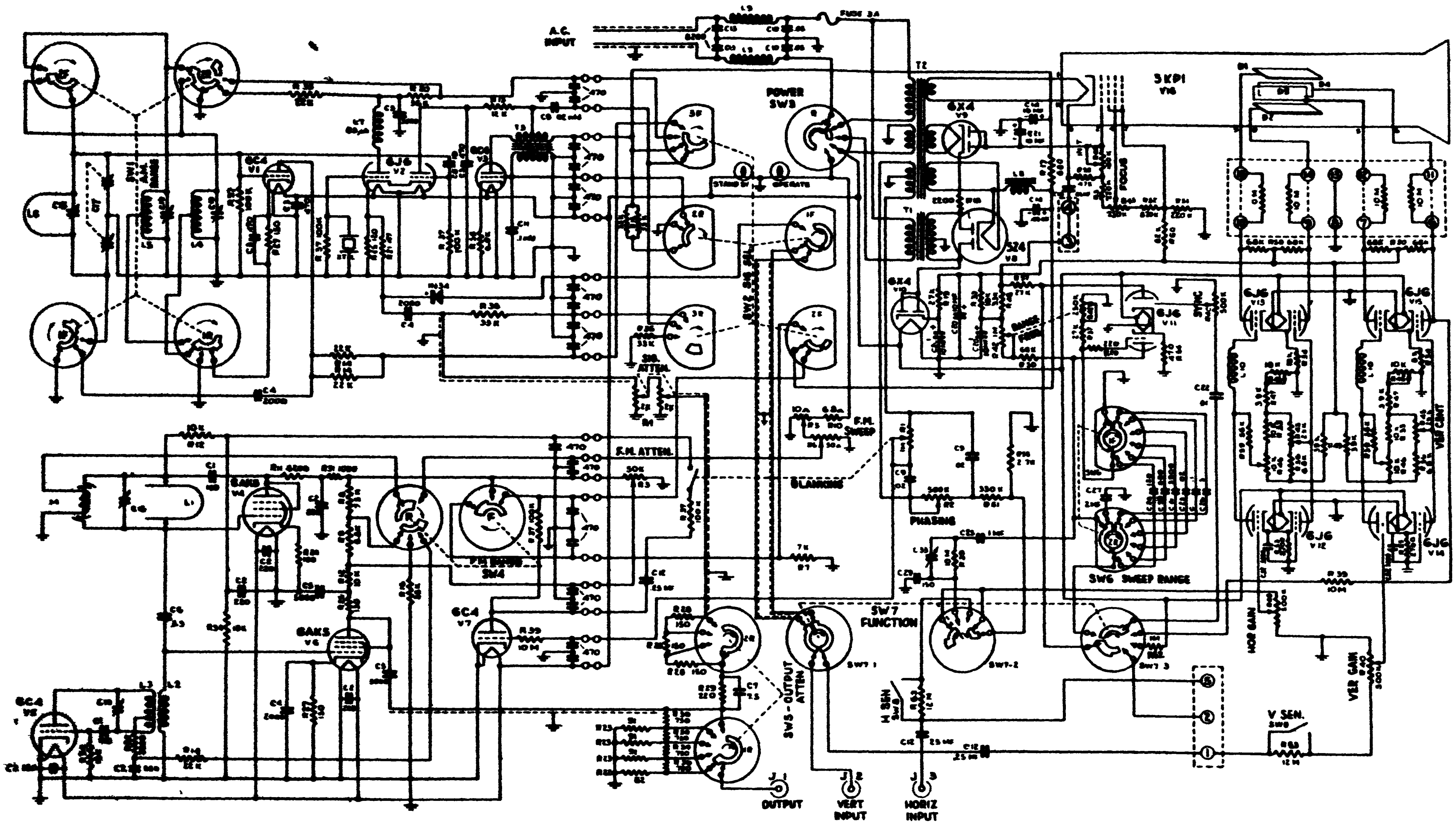
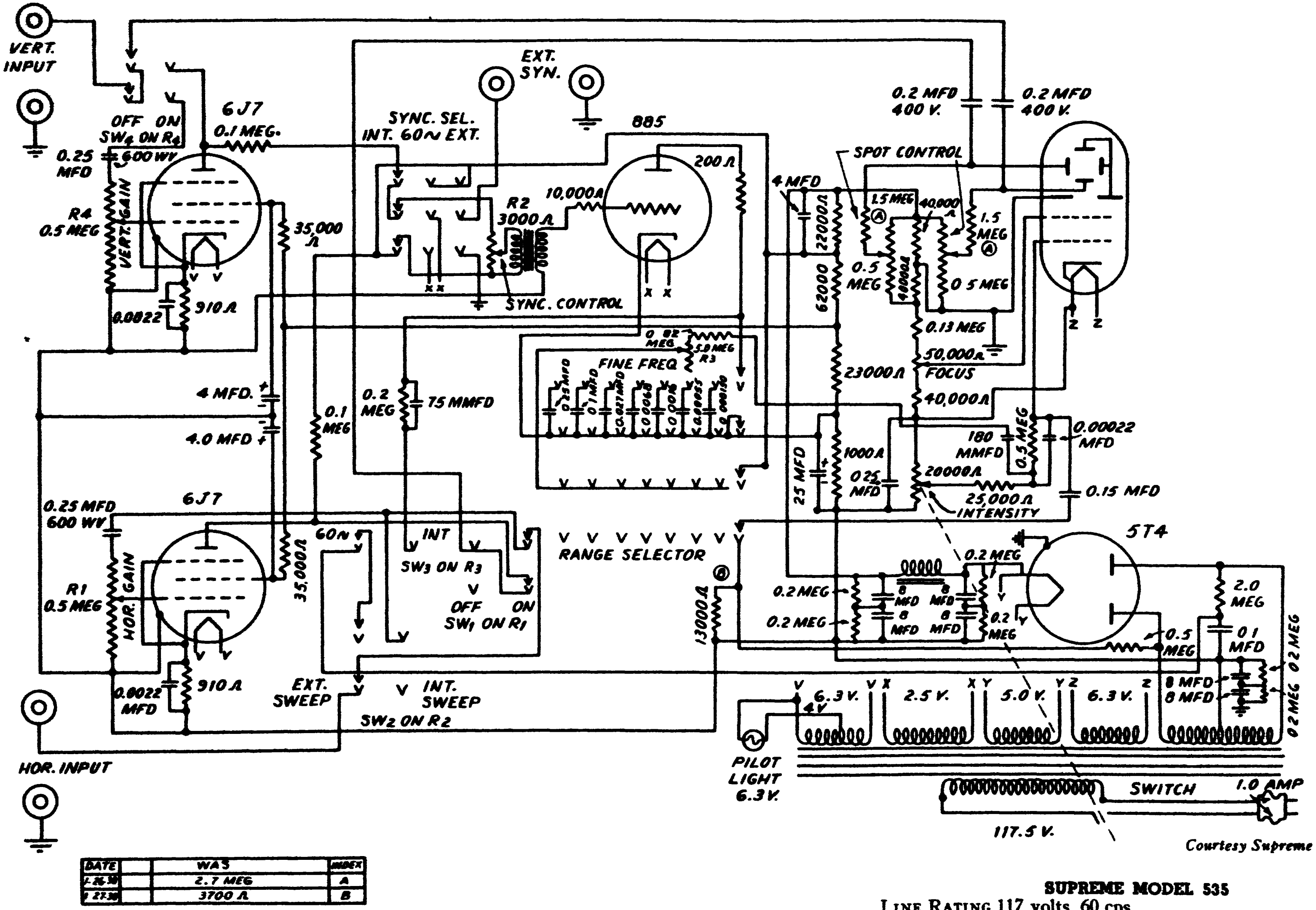


Fig. 22-60.—Schematic of Simpson Model 480.

Courtesy Simpson Elec Corp.

SIMPSON MODEL 480

Model 480, referred to as the *Genescope* and shown in Fig. 22-60, is actually a combined piece of equipment consisting of an a-m generator (marker), an f-m generator, a crystal calibrator, and an oscilloscope. Its design is principally for use in servicing and alignment of f-m and television receivers. The individual circuits themselves are conventional, but the interesting thing is their combination to perform all the desired functions. One unique feature of this unit is that the cathode-ray tube is mounted vertically. The pattern on the tube is brought into view by use of a mirror at the top of the cabinet. The mirror may be adjusted for any position of the operator.



Courtesy Supreme Inst Corp.

Fig. 22-61.—Schematic of Supreme Model 535.

SUPREME MODEL 535

LINE RATING 117 volts, 60 cps

TUBE COMPLEMENT

Type	*Function*
6J7	Vertical Amplifier
6J7	Horizontal Amplifier
885	Sweep Oscillator
5T4	Rectifier
24XH	Cathode-Ray Tube

The schematic circuit diagram of Model 535 is shown in Fig. 22-61.

SUPREME MODEL 546-A, 650, AND 655

FREQUENCY RESPONSE

Sweep Circuit 15 cps to 30 kc

LINE RATING 110-125 volts, 50-60 cps

TUBE COMPLEMENT

Type	*Function*
5CP1-A or 3AP1 (V1)	Cathode-Ray Tube
5Y3G (V2)	High-Voltage Rectifier
5Y3G (V3)	Low-Voltage Rectifier
6SJ7 (V4)	Vertical Amplifier
885 (V5)	Gaseous Sweep Oscillator
6SJ7 (V6)	Horizontal Amplifier

The circuits of Models 546-A, 650, and 655 are very much alike. The differences are chiefly in physical appearance and cathode-ray tube employed. In Fig. 22-62 appears the schematic for Model 546-A. The 546-A and 650 use a 3AP1 cathode-ray tube. Model 655 uses a 5CP1-A, a 5-inch cathode-ray tube. This necessitates use of 6.3-volt filament winding, in place of the 2.5-volt winding *B-B,* shown in Fig. 22-62. In addition, a 0.5-μf bypass capacitor has been inserted from the variable arm of *R9* to ground, in Model 655.

SUPREME MODEL 560-A

LINE RATING 117 volts, 60 cps

TUBE COMPLEMENT

Type	*Function*
1852	Input Vertical Amplifier
1852	Vertical Amplifier
1852	Vertical Amplifier
6AG7	Vertical Output
906	Cathode-Ray Tube
6SJ7	Horizontal Amplifier
884	Timing-Axis (Sweep) Oscillator
80	Low-Voltage Rectifier
80	High-Voltage Rectifier
6J5	Vacuum-Tube Voltmeter

The schematic circuit diagram of Model 560-A is shown in Fig. 22-63. This instrument is not an oscilloscope in the ordinary meaning of the word, but a combination oscilloscope, signal tracer, and electronic voltmeter. The 6J5 is used in a standard vacuum-tube-voltmeter circuit. The 1852 at the upper left may be fed audio, video, r-f, or i-f signals. A variable-capacitance attenuator links the plate circuit of this stage to the grid circuit of the succeeding 1852. This is a high-frequency stage, the plate load being only a coil.

The output of the second stage is fed to the function-selector switch. The output of this stage may be shunted to ground, or, for a proper setting of the selector switch, may feed the third 1852 stage. The third stage can also be excited from the cathode circuit of the first tube which then functions as a cathode follower, eliminating the gain of the first two 1852 stages, if desired. The plate circuit of the first 1852 may be switched to a tuned *L-C* system, for increased gain, if desired. The gain obtained is a maximum at resonance.

SUPREME MODEL 660

FREQUENCY RESPONSE

Vertical Amplifier 5 cps to 5 Mc, ±2 db
Horizontal Amplifier 5 cps to 1.5 Mc, ±2 db
Z-Axis Amplifier 100 cps to 100 kc, ±2 db
Sweep Circuit 7 cps to 150 kc

DEFLECTION FACTORS

Vertical Amplifier through probe 0.2 rms volts/inch
Vertical Amplifier 0.1 rms volts/inch
Vertical-Deflection Plates 8.5 rms volts/inch
Horizontal Amplifier 0.14 rms volts/inch
Horizontal-Deflection Plates 9 rms volts/inch

LINE RATING 110-125 volts, 50-60 cps.

The schematic circuit diagram of Model 660 is shown in Fig. 22-64. The most novel feature of this oscilloscope is its utilization of a special probe tube which permits signal tracing in the high-frequency circuits of a-m, f-m, and television sets. However, the input potential to the tube must not exceed 25 volts peak to peak. Usually, in receiver circuits, the high frequency voltages are well within the above range. The probe tube is a 6C4 which feeds into the input circuit of *V1,* the 6J5 cathode-follower vertical-input tube. The cathode follower permits the use of a low-impedance vertical-gain control *R5* which reduces frequency distortion.

V13 is a diode rectifier, not an amplifier. The signal, having passed from the probe tip *V13,* proceeds to *V1, V2, V3,* and *V4. V4* is the grounded-grid tube of the *V3-V4* push-pull vertical-output stage. This stage derives the vertical plates of the cathode-ray tube directly. There is a d-c path from each vertical plate to the corresponding plates of *V3* and *V4.*

SYLVANIA MODEL 131

FREQUENCY RESPONSE

Vertical Amplifier 10 cps to 100 kc, within 3 db
Horizontal Amplifier 10 cps to 100 kc, within 3 db
Sweep Circuit 15 cps to 40 kc

DEFLECTION FACTORS

Vertical Amplifier 0.5 rms volts/inch
Vertical-Deflection Plates 17 rms volts/inch
Horizontal Amplifier 0.5 rms volts/inch
Horizontal-Deflection Plates 17 rms volts/inch

LINE RATING 105-125 volts, 50-60 cps

TUBE COMPLEMENT

Type	*Function*
7C7 (V101)	Vertical Amplifier
7C7 (V102)	Horizontal Amplifier
884 (V103)	Sweep Oscillator
3AP1 (V104)	Cathode-Ray Tube
7Y4 (V105)	Low-Voltage Rectifier
5Y3GT (V106)	High-Voltage Rectifier

The schematic circuit diagram for Model 131 is shown in Fig. 22-65.

SYLVANIA MODEL 132

FREQUENCY RESPONSE

Vertical Amplifier 10 cps to 70 kc, ±20%
Horizontal Amplifier 10 cps to 55 kc, ±20 %
Sweep Circuit 15 cps to 30 kc

DEFLECTION FACTORS

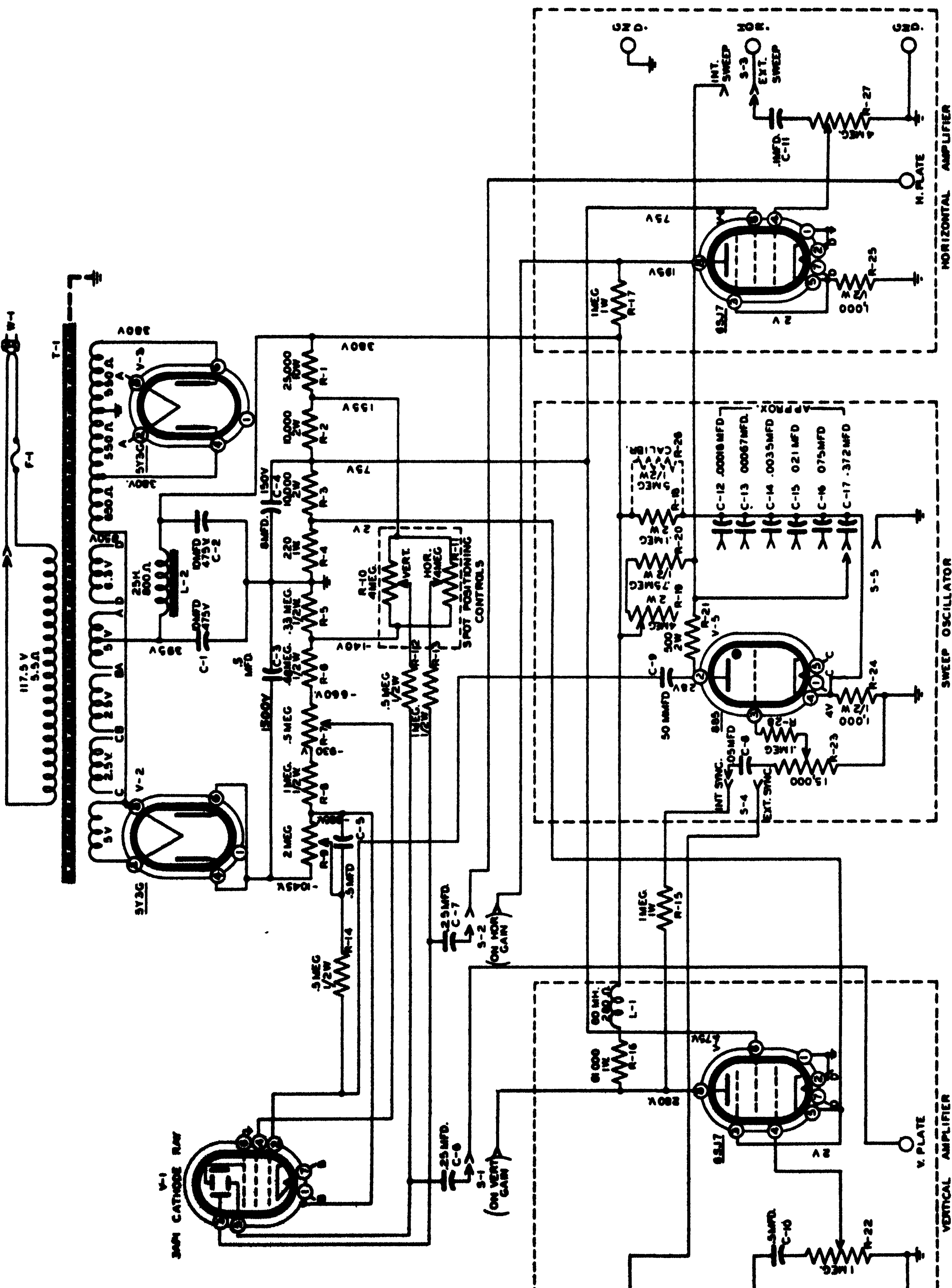
Vertical Amplifier 0.21 rms volts/inch
Vertical-Deflection Plates 15 rms volts/inch
Horizontal Amplifier 0.25 rms volts/inch
Horizontal-Deflection Plates 18 rms volts/inch

LINE RATING 105-125 volts, 50-60 cps

TUBE COMPLEMENT

Type	*Function*
7C7 (V101)	Vertical Input
7C7 (V102)	Vertical Output
7C7 (V103)	Horizontal Input
7C7 (V104)	Horizontal Output
884 (V105)	Sweep Oscillator
76P1 (V106)	Cathode-Ray Tube
2X2-A (V107)	High-Voltage Rectifier
7Y4 (108)	Full-Wave Rectifier

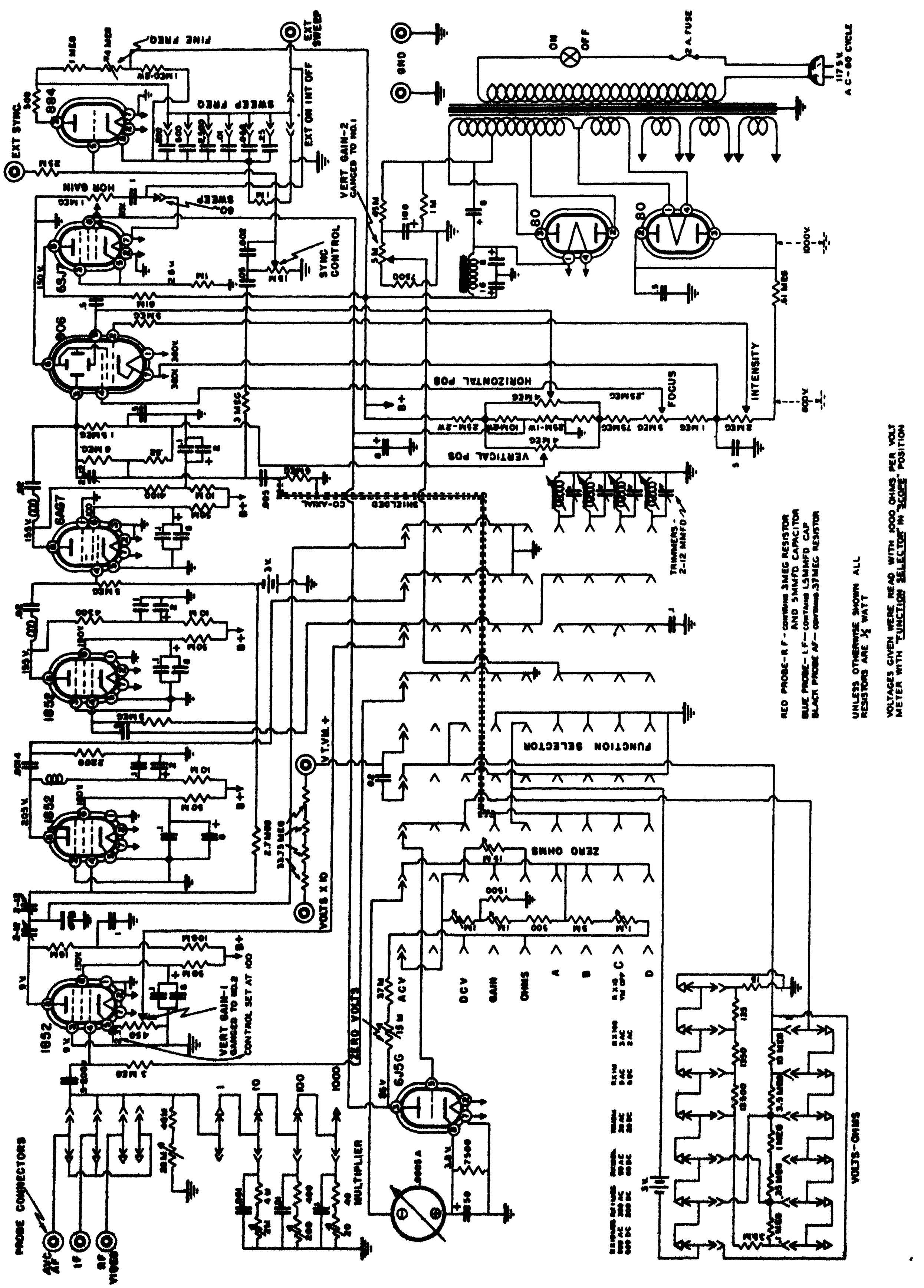
The schematic circuit diagram of Model 132 is shown in Fig. 22-66.



ALL VOLTAGES MEASURED WITH AN ELECTRONIC VOLTMETER

Fig. 22-62.—Schematic of Supreme Models 546-A, 650, 655.

Courtesy Supreme Inst. Corp.



RED PROBE-R F-CONTAINING 3MEG RESISTOR AND 5MMFD CAPACITOR
BLUE PROBE-I.F.-CONTAINING 1.5MMFD CAP
BLACK PROBE AF-CONTAINING .37MEG RESISTOR

UNLESS OTHERWISE SHOWN ALL RESISTORS ARE ½ WATT

VOLTAGES GIVEN WERE READ WITH 1000 OHMS PER VOLT METER WITH "FUNCTION SELECTOR" IN "SCOPE" POSITION

Courtesy Supreme Inst. Corp.

Fig. 22-63.—Schematic of Supreme Model 560-A.

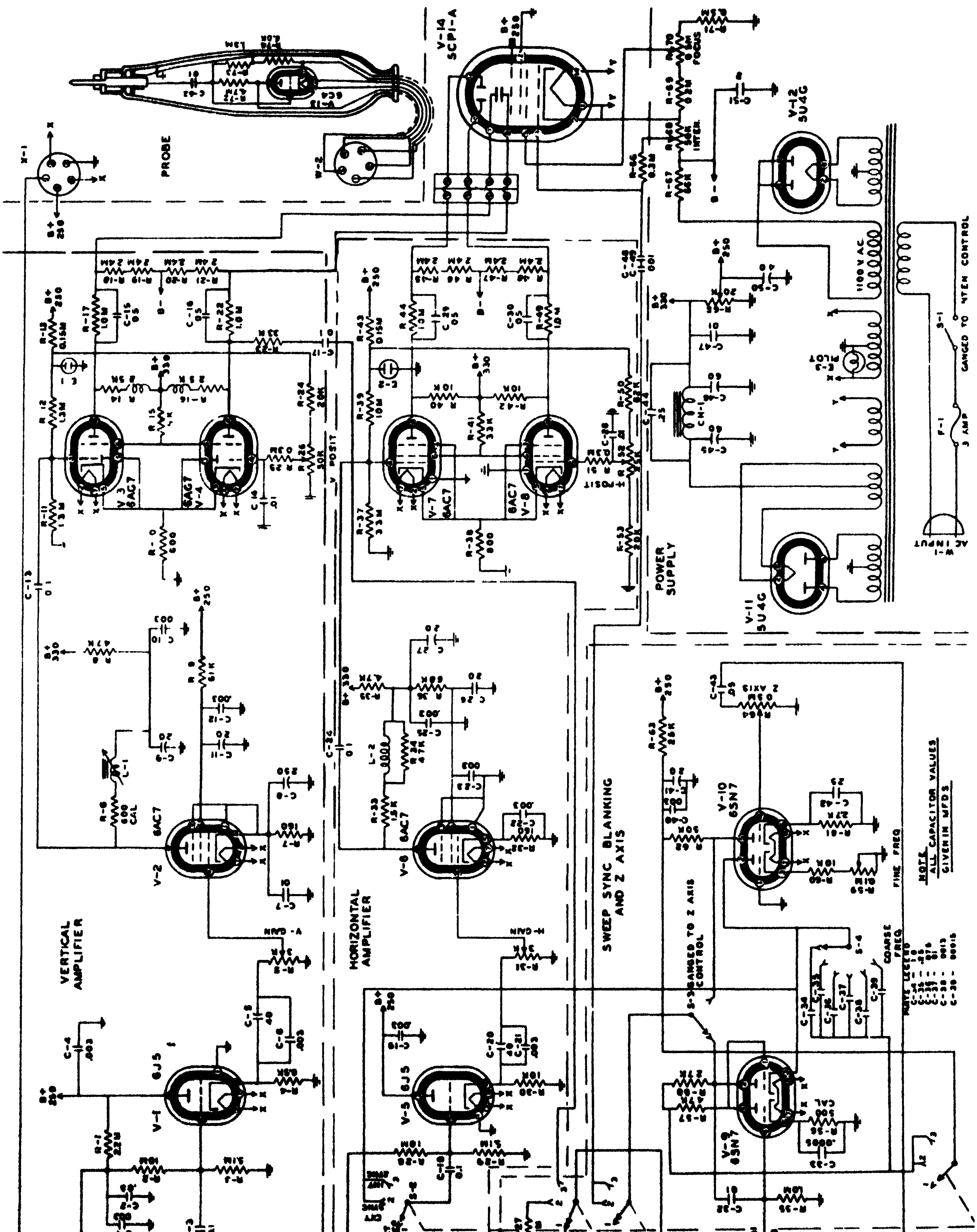
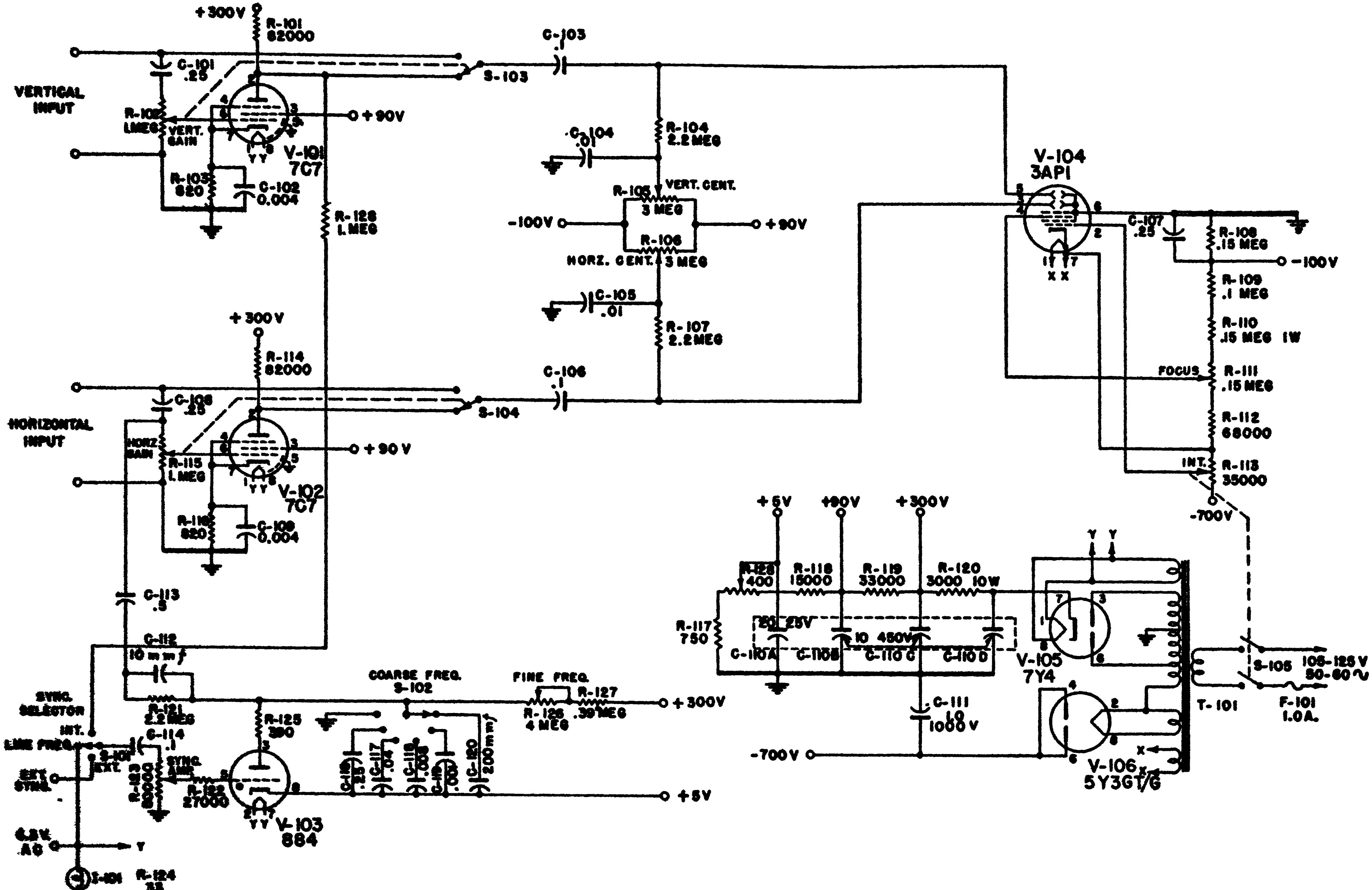


Fig. 22-64.—Schematic of Supreme Model 660

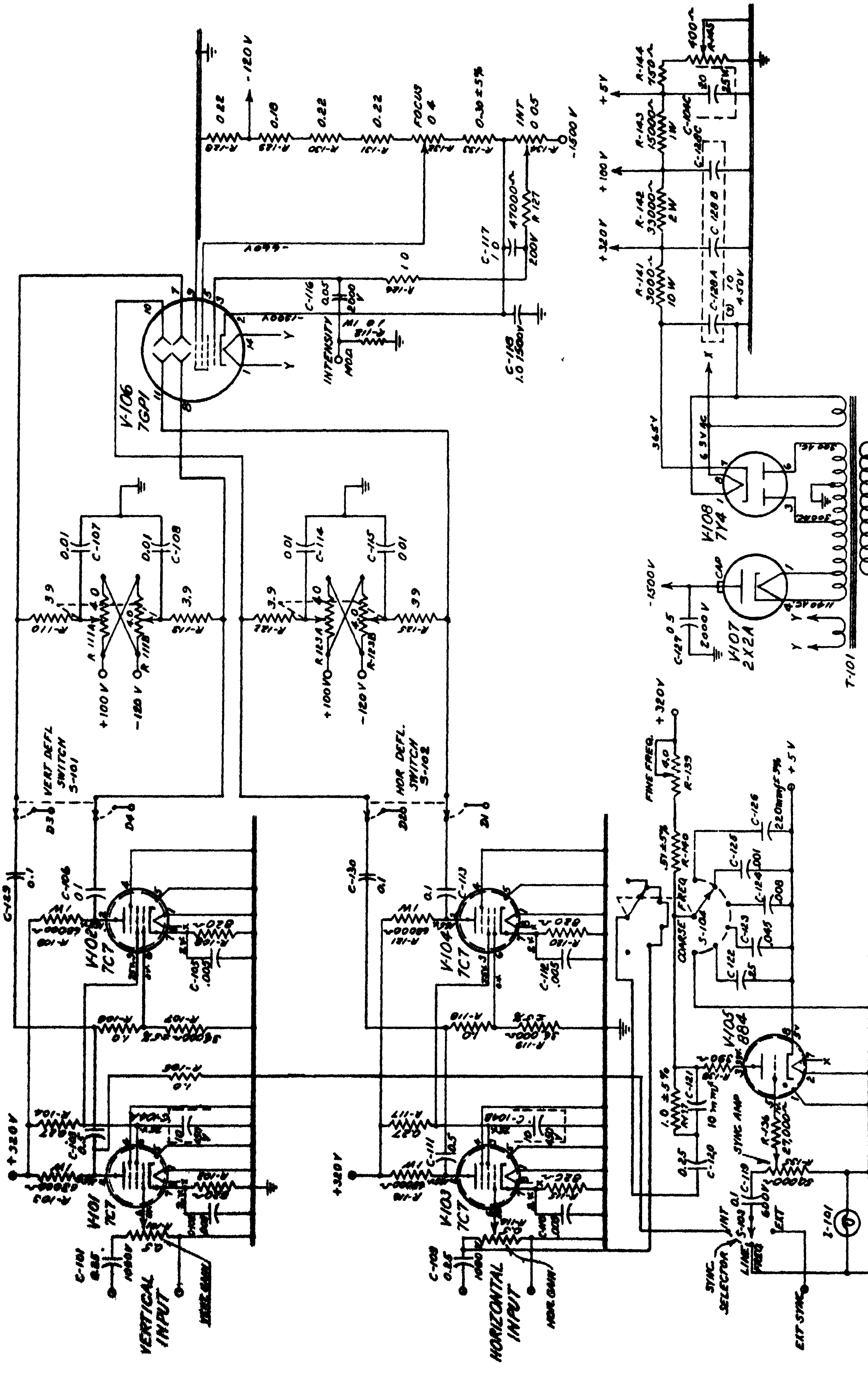
Courtesy Supreme Inst. Corp.



NOTE:—ALL CONDENSERS IN mf AND 400 VOLT RATING AND ALL RESISTORS IN OHMS AND ½ WATT UNLESS OTHERWISE NOTED

Fig. 22-63.—Schematic of Sylvania Model 131.

Courtesy Sylvania Elec. Prod.



NOTE - ALL RESISTOR VALUES IN MEGOHMS - ALL FIXED RESISTOR TOLERANCES ± 10% - ALL CAPACITOR VALUES IN mf ± 5% 400 V RATING UNLESS OTHERWISE SPECIFIED

Courtesy Sylvania Elec Prod

Fig. 22-66.—Schematic of Sylvania Model 132.

TEKTRONIX MODEL 511-A AND 511-AD

FREQUENCY RESPONSE
Vertical Amplifier 5 cps to 8 Mc, —3 db
Triggered Sweep Circuit 0.01 sec/cm to 0.1 μsec/cm
DEFLECTION FACTORS
Vertical Amplifier 0.27 rms volts/cm (minimum)
LINE RATING 105-125/210-250 volts, 50-60 cps

TUBE COMPLEMENT

Type	*Function*
6AC7 (V1)	Trigger-Phase Splitter
6AG7 (V2)	Trigger-Amplitude Control
6AL5 (V3)	Trigger-Coupling Diode
6AC7 (V4)	Multivibrator
6AG7 (V5)	Multivibrator
6AL5 (V6)	Unblanking Limiter
6C4 (V7)	Unblanking Cathode Follower
6C4 (V8)	Gate-Output Phase Splitter
6AG7 (V9)	Sweep Generator
6C4 (V10)	Sweep-Output Cathode Follower
6J6 (V11)	Sweep Magnifier
6AL5 (V12)	Sweep, D-C Restorer
6AU6 (V13)	Sweep Amplifier
6AU6 (V14)	Sweep Amplifier
6C4 (V15)	Sweep-Voltage Regulator
6AG7 (V16)	Video Amplifier, 1st Stage
6AG7 (V17-V18)	Video Amplifier, 2nd Stage
6AG7 (V19)	Video Amplifier, Gain-Control, Cathode Follower
5CP1/A (V20)	Cathode-Ray Tube
6X4 (V21)	Sweep-Supply Rectifier
5V4G (V22)	Low-Voltage Rectifier
5V4G (V23)	Low-Voltage Rectifier
6X4 (V24)	Bias Rectifier
6AQ5 (V25)	Sweep-Supply Regulator
6AU6 (V26)	Low-Voltage Regulator Amplifier
6AS7G (V27)	Low-Voltage Regulator
6AU6 (V28)	Sweep-Supply Regulator Amplifier
5651 (V29)	Voltage Reference
VR150 (V30)	Bias Regulator
6AQ5 (V31)	High-Voltage Supply Oscillator
1B3GT/8016 (V32)	High-Voltage Supply Rectifiers
1B3GT/8016 (V33)	High-Voltage Supply Rectifiers

The schematic circuit diagram for Model 511-A is shown in Figs. 22-67A and B. Model 511-AD is identical to the 511-A, except that it contains a video-delay network, operating in the cathode circuit of the *V17* video amplifier. The network provides a delay of 0.25 μsec, permitting the cathode-ray tube to be unblanked and the sweep to operate linearly, before the initiating figure reaches the vertical-deflection plates. This allows observation of random pulses.

Power Supply

The high-voltage power supply uses a unique method of developing the 3,000-volt potential required for operation of the cathode-ray tube. A 6AQ5 is used in a plate-tuned Hartley oscillator circuit, the primary of *T2* and *C75* forming the tuned circuit. At resonance, there is a large circulating current in *C75* and its associated transformer primary winding. The secondary has a low current characteristic and the 1B3GT rectifiers do not develop excessive voltage drops.

Trigger Circuit and Multivibrator

The trigger-input circuit feeds into *V1* through the switching system. *V1* has a cathode resistor and a plate-loading resistor. Across these, signal voltages are developed that may be selected by the switching system. The voltage across the series circuit of *R4* and *R5* is in phase with the input voltage on the grid, while the voltage at the plate of the phase splitter is out of phase by 180° with the grid-input voltage. *V2*, a trigger amplifier, requires a positive impulse on its grid to provide the correct trigger action for the multivibrator. When the TRIGGER SEL. switch is in the positive INT. or positive EXT. positions, signal voltages are taken off the cathode of *V1* and do not change in polarity. Therefore, a positive impulse must be supplied by the external source, or video amplifier, as the case may be.

With the TRIGGER SEL. in the negative INT. or negative EXT. positions, signals are taken off the plate, are reversed in polarity due to the phase-shifting property of the tube, and thus provide the required positive-output phase from negative-trigger sources.

To convert the various shapes of trigger impulses into square waves of controllable duration, suitable for operating the sweep generator and unblanking the cathode-ray tube, a multivibrator is provided. This circuit uses *V4* and *V5* as multivibrator tubes.

Sweep Magnifier

The function of this circuit is to delay the start of the sweep for a variable time, and then to cause it to go at five times its normal speed. This is accomplished by the biased cathode-coupled amplifier *V11*. In this type of amplifier, a positive change in the grid potential of *V11A* raises the potentials of both cathodes, and therefore corresponds to a negative change on the grid of *V11B*. Thus the plate of *V11* gives an amplified version of the signal on the grid of *V11A*, without change of polarity. The SWEEP MAGNIFIER POSITION potentiometer raises the potential on the grid of *V11B* above ground potential, and at the same time the cathode potential of *V11A*. No sweep appears on the plate of *V11B* until the plate of the sweep generator overcomes this bias. Then, the amplified sweep appears. Sufficient bias is available on the SWEEP MAGNIFIER POSITION to make it possible to start the magnified sweep at any point on the normal sweep.

Sweep Amplifier

The sweep voltage at the plate of *V9* has an amplitude of only 20 volts, approximately. To sweep the spot across the screen, about 350 volts are needed. The sweep amplifier provides the necessary amplification, and, in addition, converts the single side voltage from the sweep generator into a balanced voltage suitable for beam deflection in the cathode-ray-tube circuit. This balanced voltage is necessary to maintain the average potential of the deflection plates constant over the entire sweep range, and thus prevent defocusing. The amplifier which does this uses a cathode-coupled circuit consisting of the triode connected 6AU6 tube *V13* and *V14*.

The HOR. POSITION control varies the bias on *V14* and thus determines the position from which the sweep starts. When the SWEEP RANGE switch is on the EXT. position, the input to the sweep amplifier is shifted from the sweep generator to the arm of the EXT. SWEEP ATTEN. potentiometer.

The biasing current is obtained from the —1,500-volt cathode-ray-tube power supply through the voltage-dropping resistors *R153* and *R154*. Since the impedance of the neon-glow lamps is rather high at frequencies involved in the fastest sweeps, *C24* and *C25* are shunted across them to transmit currents at such frequencies efficiently.

Video Amplifier

A wide-range two-stage video amplifier is incorporated in the circuit, using a switching arrangement, so that only one stage need be used if desired. The first stage, which is used only in the 2 STAGES position of the INPUT CHAN. SEL. switch, uses a 6AG7 tube *V16*. The output stage is a cathode-coupled, push-pull 6AG7 stage, providing approximately 110 volts of undistorted output. Low-frequency compensation is provided to correct for the effect of the coupling capacitors. The resistors *R110* and *R125* along with the capacitors *C51A* and *C51B* provide a rising low-frequency characteristic together with a phase correction effect. To compensate for variations in the component values, *R116* is made adjustable. The inductances *L7*, *L8*, *L9* and *L10* provide high-frequency compensation. A four-terminal coupling network is used to provide the greatest possible bandwidth. The 6AG7 cathode follower *V17* serves primarily as an impedance transformer, so that a low-resistance potentiometer may be used as a gain control. The cathode follower permits the insertion of a delay line in the video amplifier, which is provided for Model 511-AD.

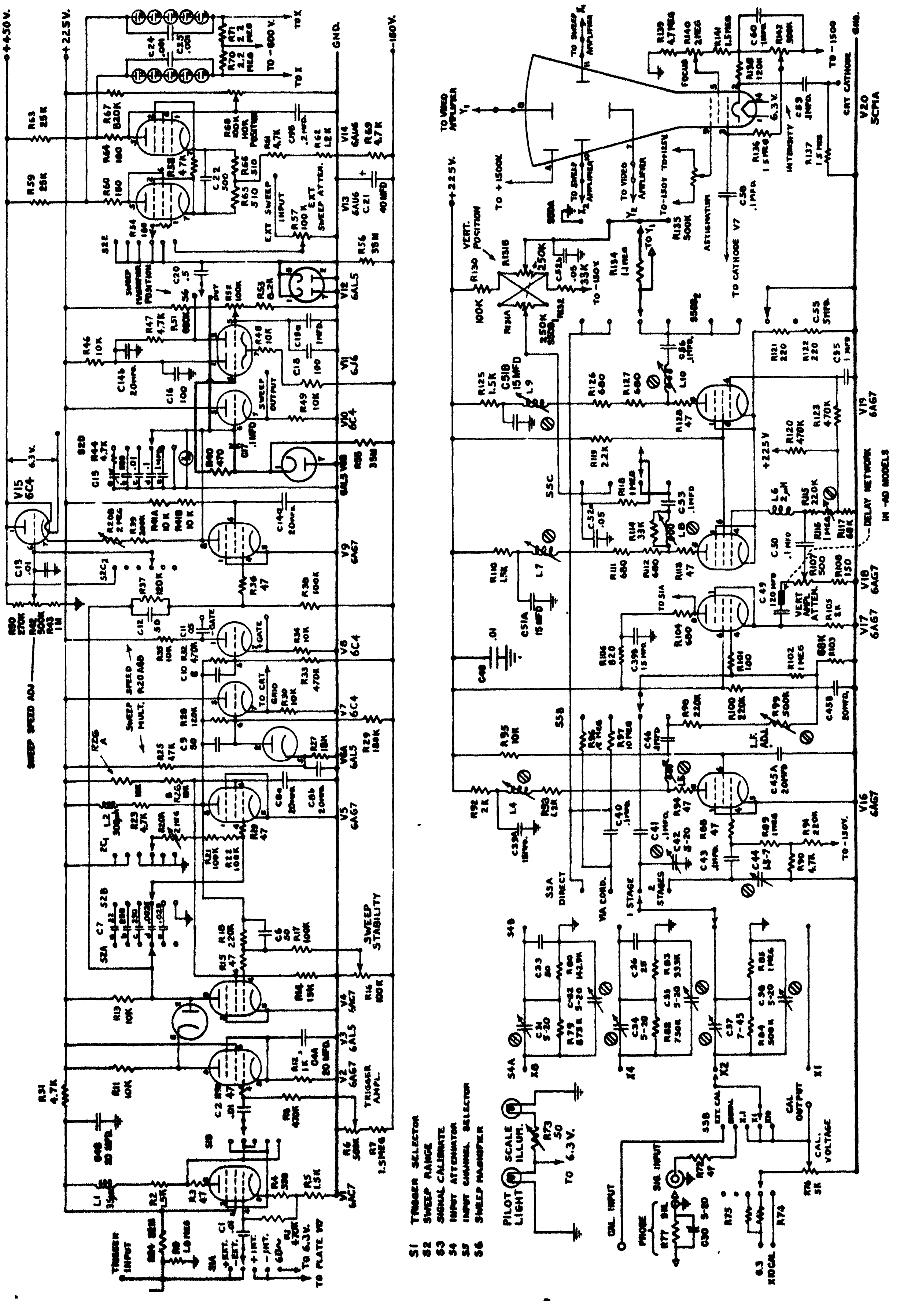


Fig. 22-67A.—Schematic of Tektronix Model 511-A.

Courtesy Tektronix Inc.

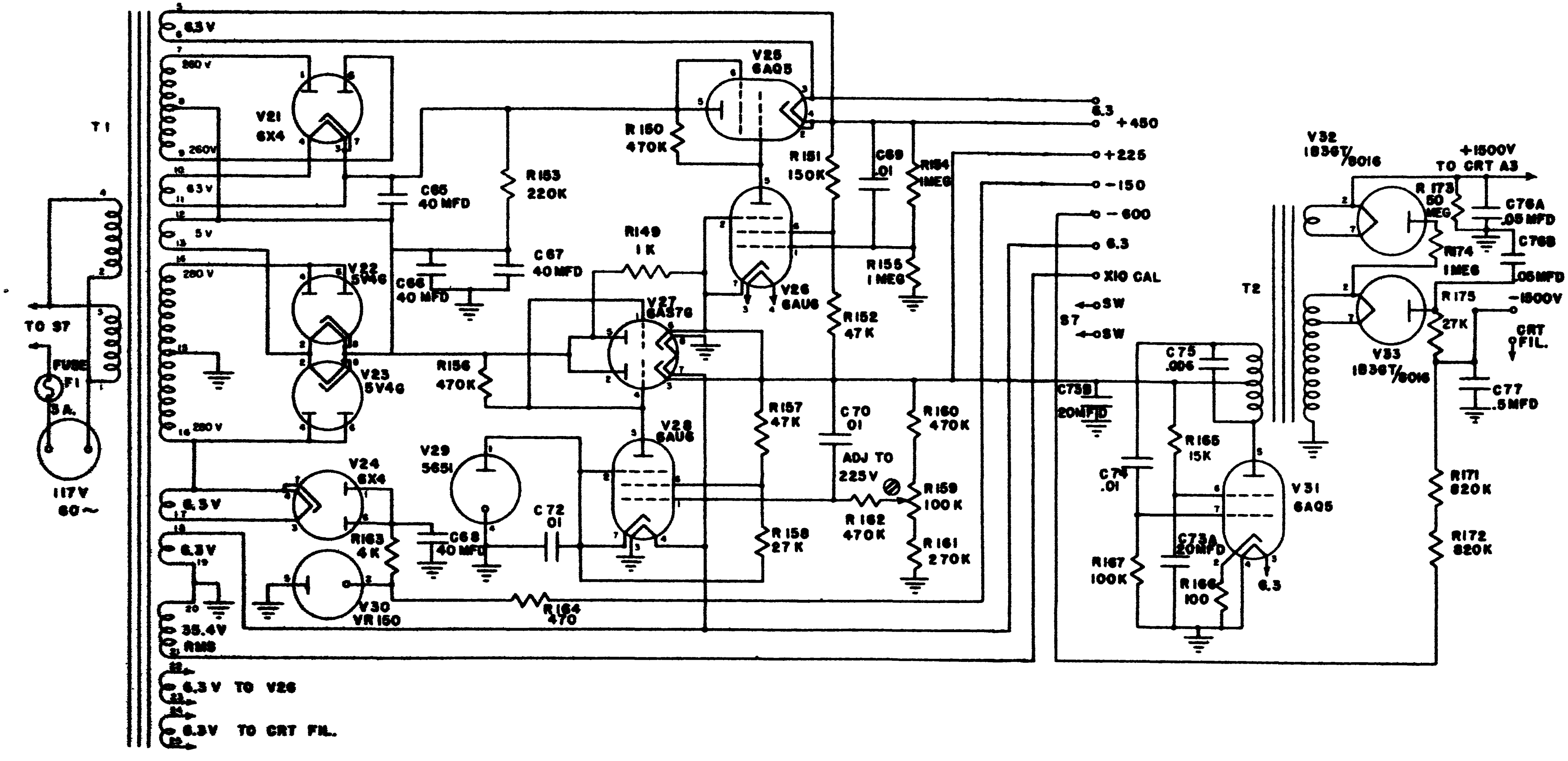


Fig. 22-67B.—Power supply of Tektronix Model 511-A.

Courtesy Tektronix Inc

TELEVISION EQUIPMENT CORP. MODEL T-602

Frequency Response

- Vertical Amplifier 2 cps to 750 kc, ±10%
- Horizontal Amplifier 2 cps to 325 kc, ±10%
- Sweep Circuit 1 cps to 50 kc

Deflection Factors

- Vertical Amplifier 1 rms mv/inch
- Vertical-Deflection Plates 215 rms volts/inch
- Horizontal Amplifier 60 rms mv/inch
- Z-Axis Amplifier 2 volts rms visible

The schematic circuit diagram of Model T-602 is shown in Fig. 22-68.

TRIUMPH MODEL 840

Frequency Response

- Vertical Amplifier 10 cps to 100 kc
- Horizontal Amplifier 10 cps to 100 kc
- Sweep Circuit 7 cps to 30 kc

Deflection Factors

- Vertical Amplifier 0.4 rms volts/inch
- Vertical-Deflection Plates 18 rms volts/inch
- Horizontal Amplifier 0.45 rms volts/inch
- Horizontal-Deflection Plates 20 rms volts/inch

Line Rating 115 volts, 50-60 cps

Tube Complement

Type	*Function*
6SJ7 (VT1)	Vertical Amplifier
6SJ7 (VT2)	Horizontal Amplifier
884 (VT3)	Sweep Oscillator
6J5 (VT4)	Return Trace-Blanking Tube
3AP1/906P1 (VT5)	Cathode-Ray Tube
1V (VT6)	High-Voltage Rectifier
6X5 (VT7)	Low-Voltage Rectifier

The schematic circuit diagram for Model 840 is shown in Fig. 22-69.

TRIUMPH MODEL 841

Frequency Response

- Vertical Amplifier From 10 cps to 100 kc, ±2 db
- Horizontal Amplifier From 10 cps to 100 kc, ±2 db
- Sweep Circuit 7 cps to 30 kc

Deflection Factors

- Vertical Amplifier 0.8 rms volts/inch
- Vertical-Deflection Plates 31.5 rms volts/inch
- Horizontal Amplifier 1 rms volts/inch
- Horizontal-Deflection Plates 42.5 rms volts/inch

Line Rating 105-125 volts, 50-70 cps

Tube Complement

Type	*Function*
6AC7 (V101)	Vertical Amplifier
6AC7 (V102)	Horizontal Amplifier
884 (V103)	Sweep Oscillator
3BP1 (V104)	Cathode-Ray Tube
2X2 (V105)	High-Voltage Rectifier
6X5GT (V106)	Low-Voltage Rectifier

The schematic circuit diagram of Model 841 is shown in Fig. 22-70.

TRIUMPH MODEL 850

Frequency Response

- Vertical Amplifier 5 cps to 50 kc
- Horizontal Amplifier 5 cps to 100 kc
- Sweep Circuit 5 cps to 30 kc

Deflection Factors

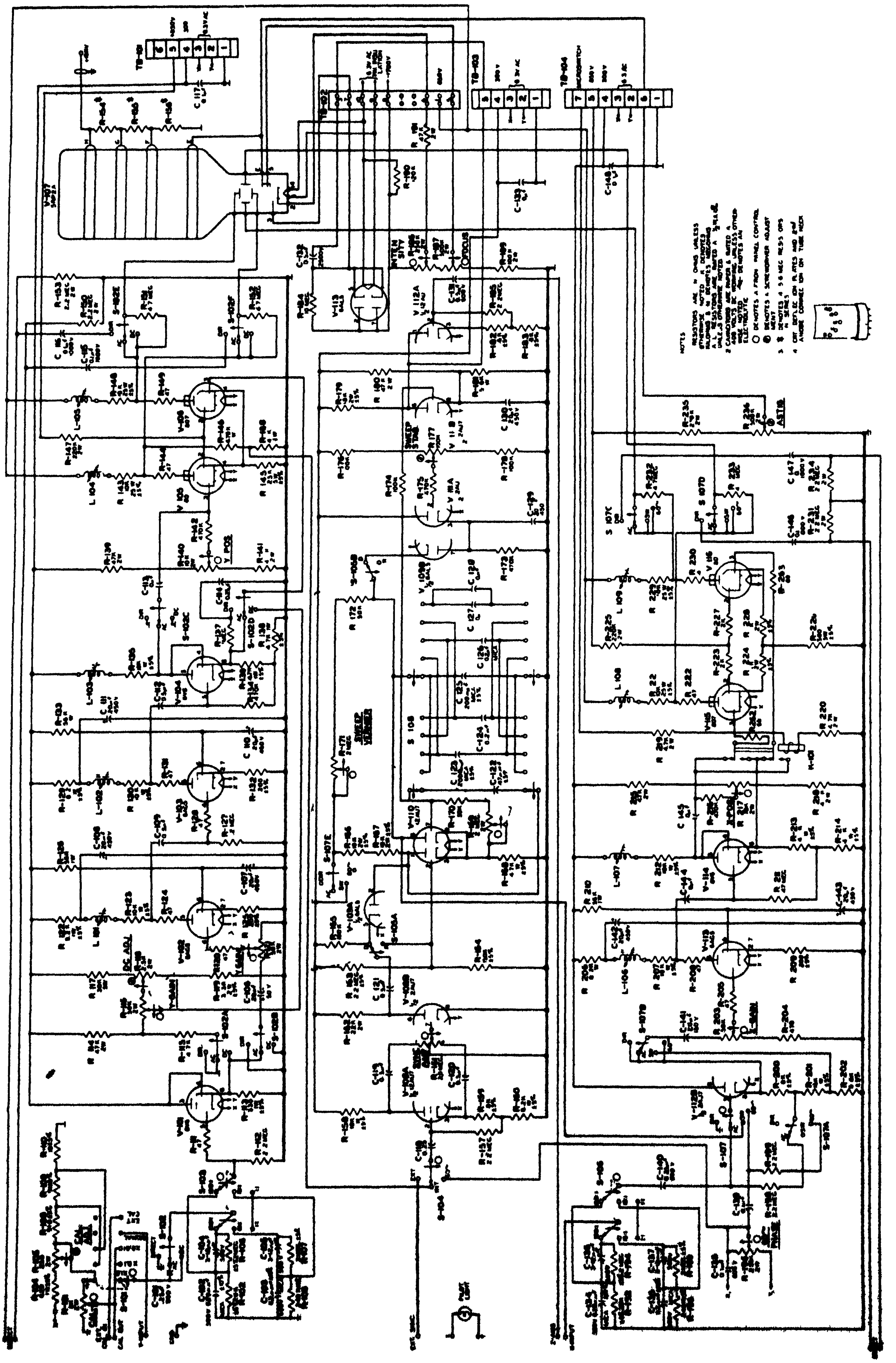
- Vertical Amplifier 0.02 rms volts/inch
- Vertical-Deflection Plates Direct 20 rms volts/inch
- Horizontal Amplifier 0.45 rms volts/inch
- Horizontal-Deflection Plates Direct 20 rms volts/inch

Line Rating 115 volts, 40-60 cps

Tube Complement

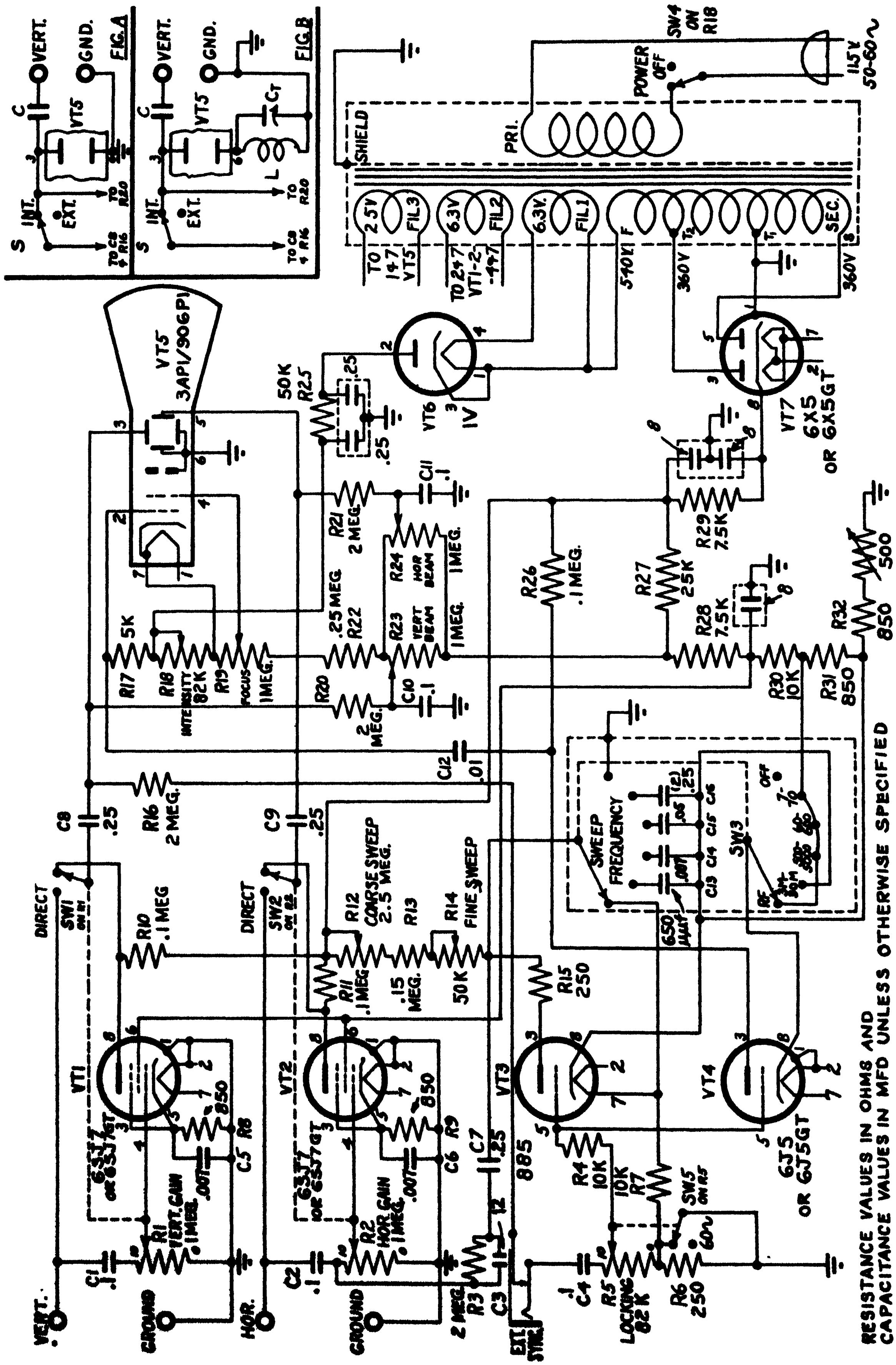
Type	*Function*
6SJ7 (VT1)	Vertical-Amplifier Input
6SJ7 (VT2)	Vertical-Amplifier Output
6SJ7 (VT3)	Horizontal Amplifier
884 (VT4)	Sweep Oscillator
6J5 (VT5)	Blanking Tube
5BP1/1802P1 (VT6)	Cathode-Ray Tube
2X2/879 (VT7)	High-Voltage Rectifier
6X5 (VT8)	Low-Voltage Rectifier

The schematic circuit diagram of Model 850 is shown in Fig. 22-71.



Courtesy Television Equip Corp

Fig. 22-68.—Schematic of TEC Model T-602.



RESISTANCE VALUES IN OHMS AND CAPACITANCE VALUES IN MFD UNLESS OTHERWISE SPECIFIED

Fig. 22-69.—Schematic of Triumph Model 840.

Courtesy Triumph Mfg. Co.

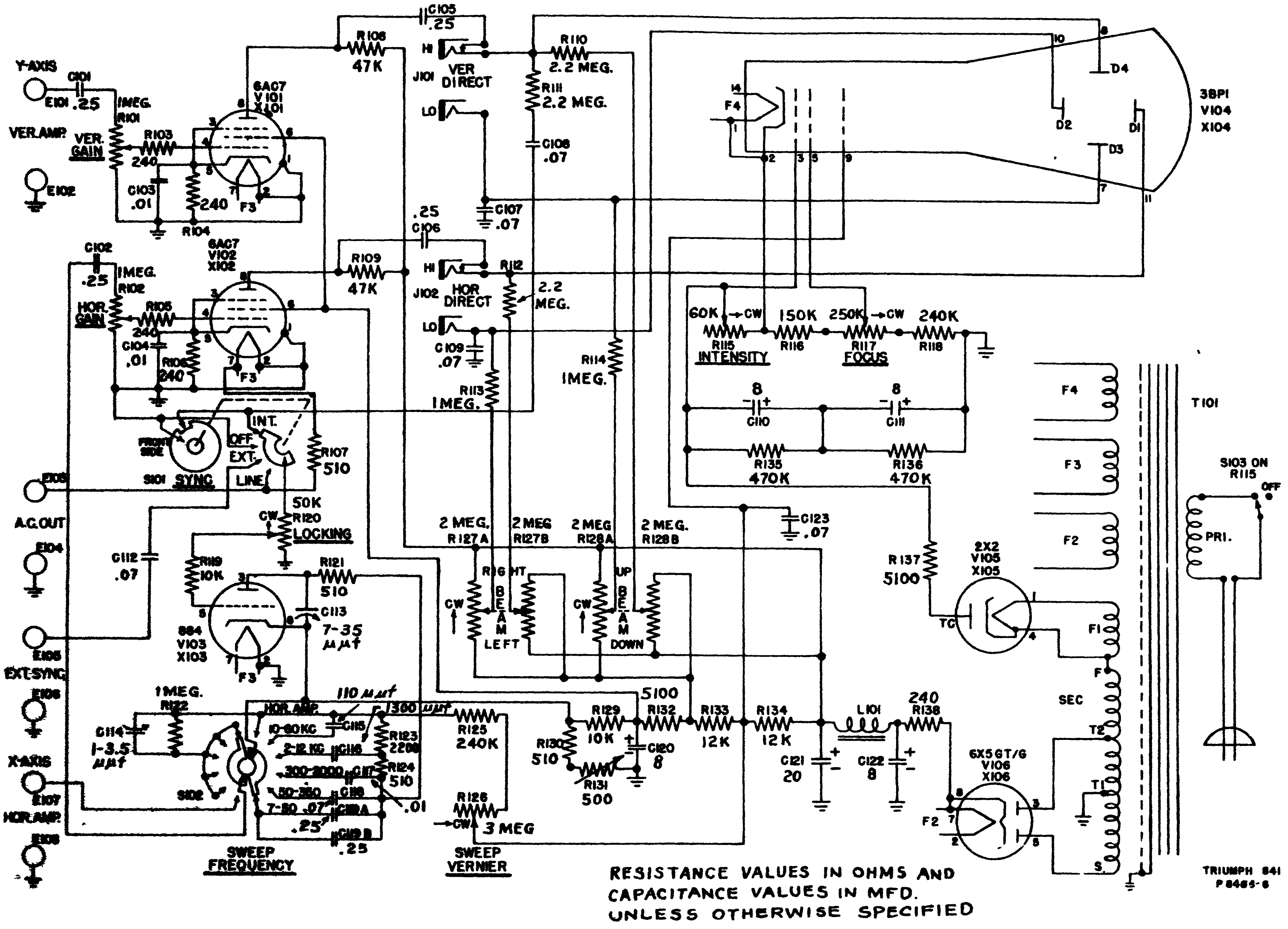
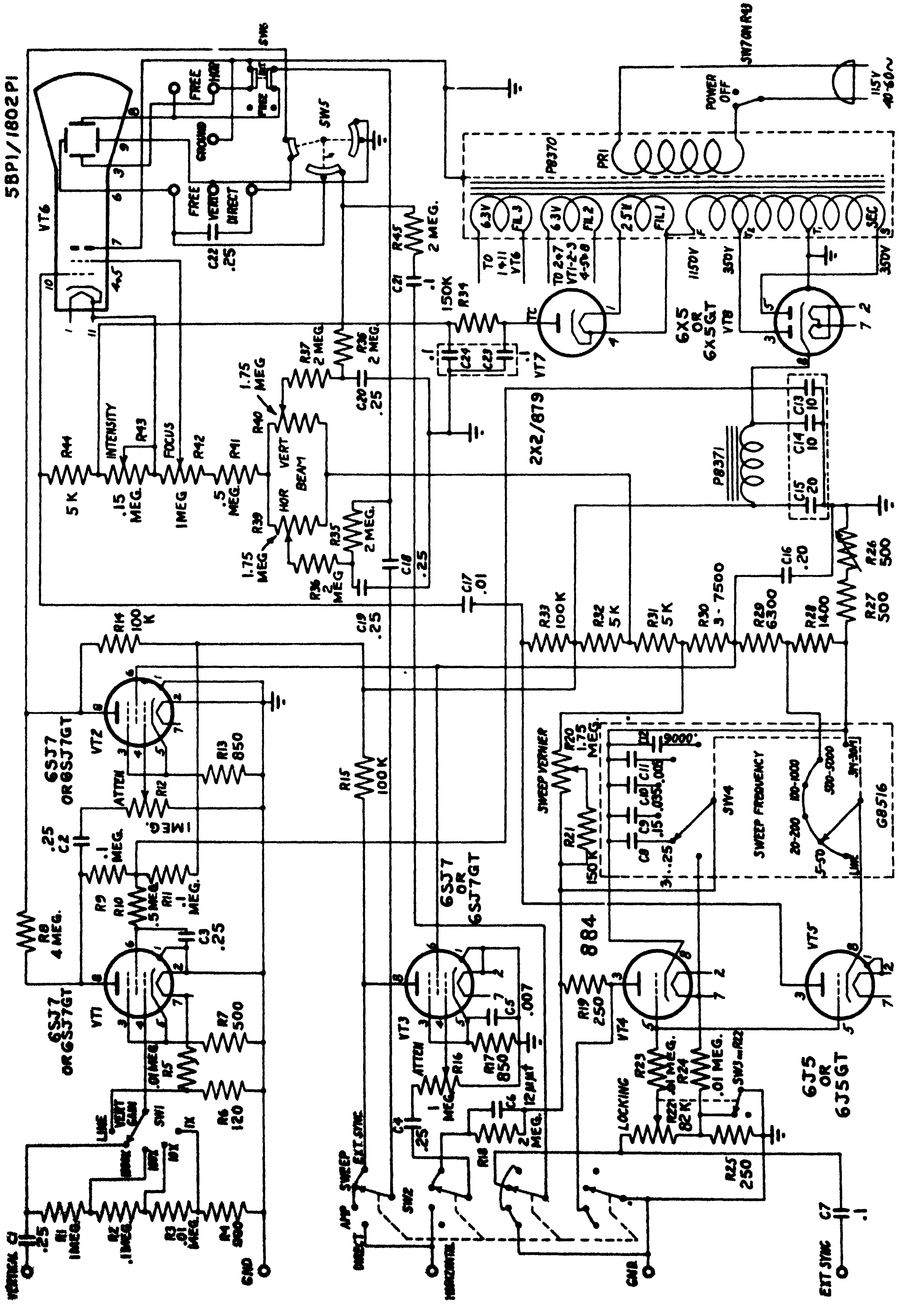


Fig. 22-70.—Schematic of Triumph Model 841.

Courtesy Triumph Mfg. Co.



RESISTANCE VALUES IN OHMS AND CAPACITANCE VALUES IN MFD. UNLESS OTHERWISE SPECIFIED

Fig. 22-71.—Schematic of Triumph Model 850.

Courtesy Triumph Mfg Co

WATERMAN MODEL S-10-A

FREQUENCY RESPONSE
  Vertical Amplifier 20 cps to 100 kc, —2 db
  Horizontal Amplifier 20 cps to 100 kc, —2 db
  Sweep Circuit 10 cps to 50 kc
DEFLECTION FACTORS
  Vertical Amplifier 1 rms volts/inch
  Vertical-Deflecton Plates 30 rms volts/inch
  Horizontal Amplifier 1 rms volts/inch
  Horizontal-Deflection Plates 24 rms volts/inch
LINE RATING 105-125 volts, 50-60 cps

TUBE COMPLEMENT

Type	*Function*
6AU6 (X1)	Vertical Amplifier
2AP1-A (X2)	Cathode-Ray Tube
6J6 (X3)	Sweep Oscillator and Horizontal Amplifier
6X4 (X4)	Rectifier

The schematic circuit diagram of Model S-10-A is shown in Fig. 22-72.

WATERMAN MODEL S-10-B

FREQUENCY RESPONSE
  Vertical Amplifier 20 cps to 150 kc, —2 db
  Horizontal Amplifier 20 cps to 150 kc, —2 db
  Sweep Circuit 10 cps to 50 kc
DEFLECTION FACTORS
  Vertical Amplifier 1 rms volts/inch
  Vertical-Deflection Plates 30 rms volts/inch
  Horizontal Amplifier 1 rms volts/inch
  Horizontal-Deflection Plates 30 rms volts/inch
LINE RATING 105-125 volts, 50/60 cps

TUBE COMPLEMENT

Type	*Function*
12AT7 (V1)	Vertical and Horizontal Amplifier
6J6 (V2)	Time-Base Generator
117Z6GT (V3)	High- and Low-Voltage Supply Rectifier
2AP1A (V4)	Cathode-Ray Tube

The schematic circuit diagram for Model S-10-B is shown in Fig. 22-73.

WATERMAN MODEL S-11-A

FREQUENCY RESPONSE
  Vertical Amplifier 0 cps to 200 kc, —2 db
  Horizontal Amplifier 0 cps to 200 kc, —2 db
  Sweep Circuit 3 cps to 50 kc
DEFLECTION FACTORS
  Vertical Amplifier 0.1 rms volts/inch
  Vertical-Deflection Plates 28 rms volts/inch
  Horizontal Amplifier 0.1 rms volts/inch
  Horizontal-Deflection lates 28 rms volts/inch
LINE RATING 105-125 volts, 50/60 cps

TUBE COMPLEMENT

Type	*Function*
6J6 (V1)	Vertical-Input Amplifier
6J6 (V2)	Vertical-Output Amplifier
6J6 (V3)	Horizontal-Input Amplifier
6J6 (V4)	Horizontal-Output Amplifier
6J6 (V5)	Time-Base Oscillator, Blanking and Intensity Amplifier
117Z6/GT (V6)	Low- and High-Voltage Rectifier
3MP1 (V7)	Cathode-Ray Tube

The schematic circuit diagram for Model S-11-A is shown in Fig. 22-74. The instrument uses balanced directly coupled vertical and horizontal amplifiers consisting of dual-triode 6J6's to assure wide frequency response.

Intensity Modulation

Upon rotation of the FUNCTION switch *S2* to HOR, the horizontal-input posts are connected to the horizontal amplifier, the linear time-base generator is made inoperative, and *V5* becomes an intensity amplifier for modulating the intensity of the beam. The plate of the intensity amplifier *V5* is coupled to the cathode of the cathode-ray tube; thus a positive signal will produce intensification of the beam.

WATERMAN MODEL S-12-A

FREQUENCY RESPONSE
  Vertical Amplifier dc to 200 kc, —2 db
  Horizontal Amplifier dc to 200 kc, —2 db
  Sweep Circuit 0.5 cps to 50 kc
DEFLECTION FACTORS
  Vertical Amplifier 0.05 rms volts/inch
  Vertical-Deflection Plates 30 rms volts/inch
  Horizontal Amplifier 0.05 rms volts/inch
  Horizontal-Deflection Plates 30 rms volts/inch
LINE RATING 105-125 volts, 50-60 cps

TUBE COMPLEMENT

Type	*Function*
12AT7 (V1)	Vertical-Input Amplifier
12AU7 (V2)	Vertical Cathode Amplifier
12AT7 (V3)	Vertical-Output Amplifier
12AT7 (V4)	Horizontal-Input Amplifier
12AU7 (V5)	Horizontal Cathode Follower
12AT7 (V6)	Horizontal-Output Amplifier
12AU7 (V7)	Sync Polarity
12AT7 (V9)	Sweep Generator
12AT7 (V10)	Sweep Generator
6X4 (11)	Positive-Voltage Supply
6X4 (V12)	Negative-Voltage Supply
12AX7 (V13)	Intensity-Modulation Amplifier
3MP1 (V14)	Cathode-Ray Tube

The schematic circuit diagram for Model S-12-A is shown in Fig. 22-75. The intensity-modulation control permits blanking of the return trace in the extreme counterclockwise position. When the control is rotated clockwise, it ties the input to the intensity binding post to the control grid of the intensity modulation amplifier *V13*. Hence, a signal applied to the INT binding post causes modulation of the cathode-ray-tube beam intensity.

Tube *V2* is a duo-triode cathode follower which serves to isolate the output capacitance of the first stage from the input capacitance of the third stage. This action aids in the extension of the over-all frequency range of the amplifier. The third tube, a duo-triode 12AT7 *V3* functions as a phase inverter and push-pull output stage.

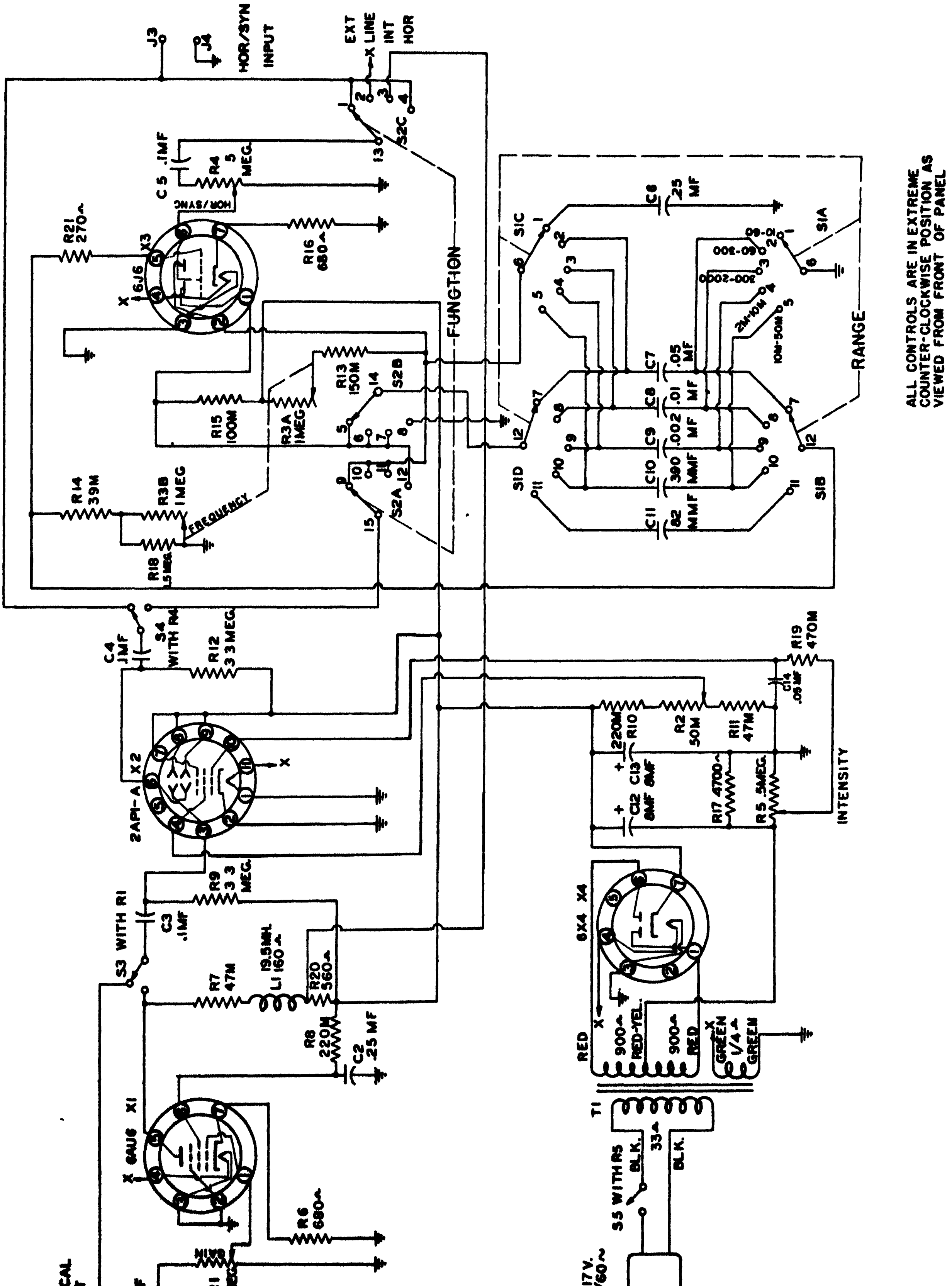
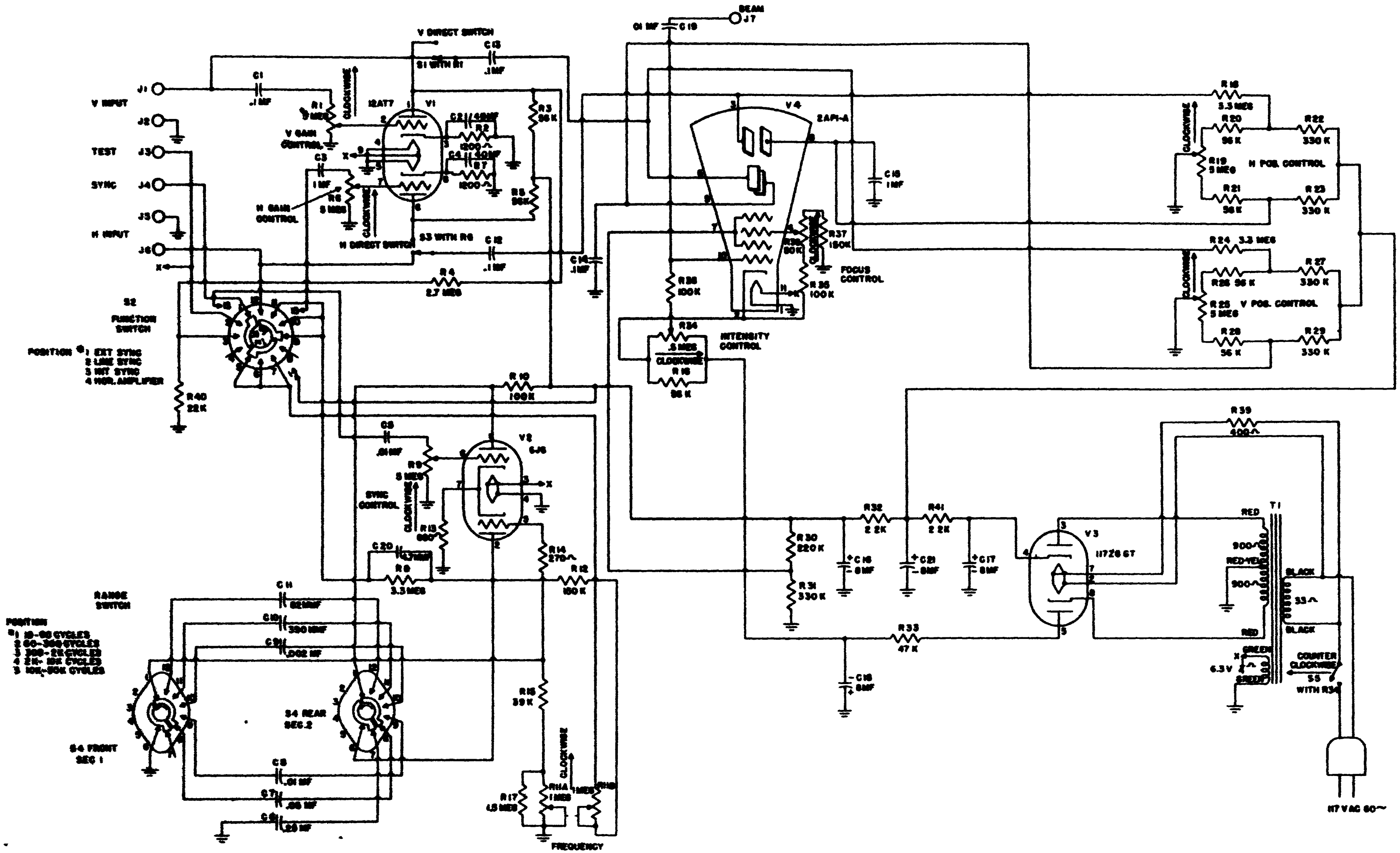


Fig. 22-72.—Schematic of Waterman Model S-10-A.

Courtesy Waterman Prod Co.



NOTE ALL SWITCH SECTIONS VIEWED FROM REAR
FRONT OF SECTION #1 IS NEAREST PANEL
SWITCHES SHOWN IN #1 POSITION
DIRECTION OF ROTATION IS SHOWN BY ARROWS
SWITCHES WHEN VIEWED FROM FRONT
ARE IN EXTREME COUNTER-CLOCKWISE
POSITION & ROTATION IS IN CLOCKWISE DIRECTION

Fig. 22-73.—Schematic of Waterman Model S-10-B.

Courtesy Waterman Prod. Co.

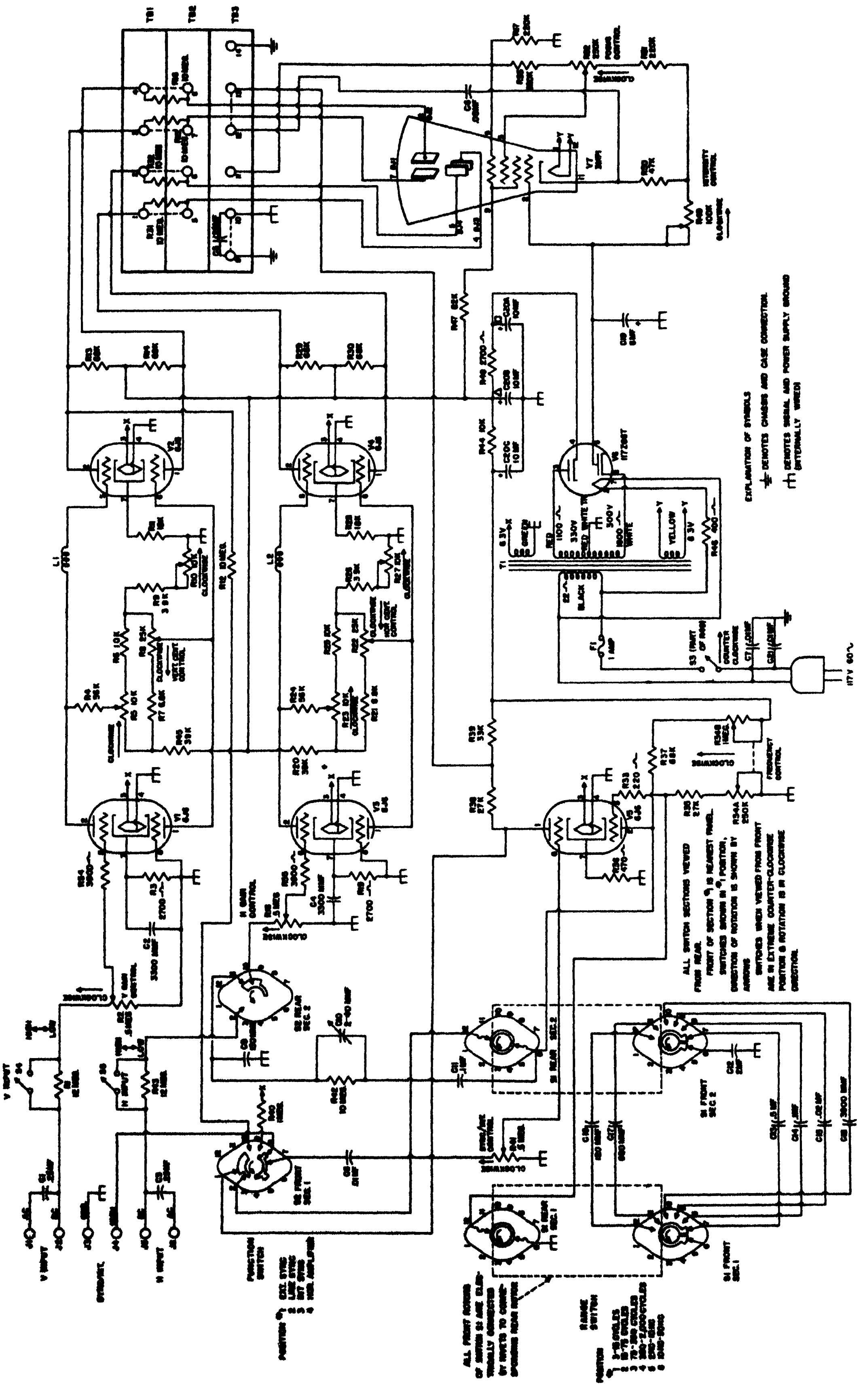
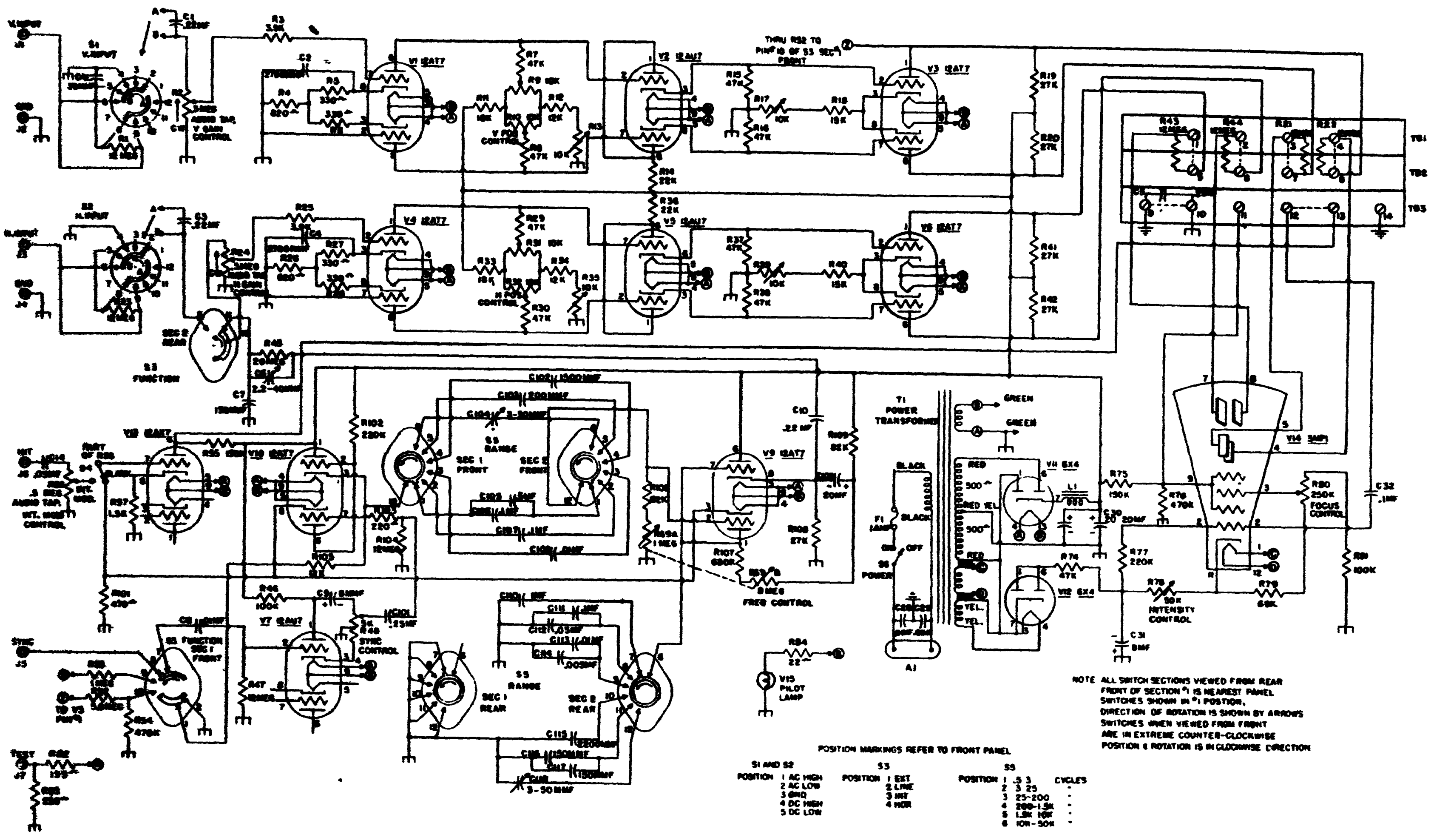


Fig 22-74—Schematic of Waterman Model S-11-A.

Courtesy Waterman Prod Co



NOTE ALL SWITCH SECTIONS VIEWED FROM REAR
FRONT OF SECTION #1 IS NEAREST PANEL
SWITCHES SHOWN IN #1 POSITION,
DIRECTION OF ROTATION IS SHOWN BY ARROWS
SWITCHES WHEN VIEWED FROM FRONT
ARE IN EXTREME COUNTER-CLOCKWISE
POSITION & ROTATION IS IN CLOCKWISE DIRECTION

POSITION MARKINGS REFER TO FRONT PANEL

S1 AND S2
POSITION 1 AC HIGH
2 AC LOW
3 GND
4 DC HIGH
5 DC LOW

S3
POSITION 1 EXT
2 LINE
3 INT
4 HOR

S5
POSITION 1 .5 3 CYCLES
2 3 25 "
3 25-200 "
4 200-1.5K "
5 1.5K 10K "
6 10K-50K "

Fig. 22-75.—Schematic of Waterman Model S-12-A.

Courtesy Waterman Prod. Co.

APPENDIX I

CHARACTERISTICS OF CATHODE-RAY TUBES

ELECTROSTATIC TYPES—CATHODE RAY TUBES

Type	Heater		Nominal Dimensions		Base	RMA Basing	Screen		Maximum Design Center Ratings				Typical Operating Conditions				Deflection Avg. Volts DC/Inch	
	Volts	Amperes	Diameter Inches	Length Inches			Fluorescence	Persistence	Anode #1 Volts	Anode #2 Volts	Anode #3 Volts	Anode #2 to Deflection Plate Peak Volts	Anode #2 Volts	Anode #1 Avg. Volts**	Anode #3 Volts	Grid Range Volts*	D 1-2	D 3-4
2AP1) 2AP1A)	6.3	0.6	2	7-7/16	Small Shell Magnal 11 Pin	11B 11L	Green	Medium	500	1000	. .	600	500 1000	125 250	....	15-45 30-90	115 230	98 196
2BP1	6.3	0.6	2-1/16	7-5/8	Small Shell Duodecal 12 Pin	12E 12E	Green	Medium	1000	2500	...	500	1000 2000	150-280 300-560	....	0-67.5 0-135	115-155 230-310	74-100 148-200
3AP1) 3AP1A) 3AP4)	2.5	2.1	3	11-1/2	Medium 7 Pin	7AN 7CE 7AN	Green Green White	Medium Medium Medium	1000	1500	. ..	600	600 800 1000 1200 1500	170 230 285 345 475	....	14-40 14-40 17-50 20-60 22.5-67.5	47 61 76 91 114	45 58 73 87 109
3BP1) 3BP1A)	6.3	0.6	3	10	Medium Shell Diheptal 12 Pin	14A 14A	Green Green	Medium Medium	1000	2000	...	500	1500 2000	430 575	....	22.5-67.5 30-90	168 221	123 164
3CP1	6.3	0.6	3	10-3/8	Medium Magnal 11 Pin, Sleeve	11C	Green	Medium	1000	2000	....	500	1500 2000	430 575	....	22.5-67.5 30-90	165.5 124	221 165
3DP1) 3DP1A)	6.3	0.6	3	10-7/16	Medium Shell Diheptal 12 Pin	14C 14H	Green	Medium	1000	2000	...	500	1500 2000	430 575	....	22.5-67.5 30-90	166 221	123 164
3EP1	6.3	0.6	3	9-15/16	Large Wafer Magnal 11 Pin, Sleeve	11A	Green	Medium	1000	2000	....	500	1500 2000	430 575	....	22.5-67.5 30-90	165.5 221	124 165
3FP7 3FP7A	6.3	0.6	3	10	Medium Shell Diheptal 12 Pin	14B 14J	See chap.7 for Characteristics of Phosphor No 7		1000	2000	4000	500	2000 1500 2000	575 430 575	2000 3000 4000	30-90 22.5-67.5 30-90	221 221 295	164 163 217
3GP1 3GP4	6.3	0.6	3	11-1/2	Medium Shell Magnal 11 Pin	11A 11A	Green White	Medium Medium	1000	1500	. .	500	1000 1500	234 350	....	16.5-49.5 25-75	80 120	70 105
3GP1A 3GP4A	6.3	0.6	3	11-1/2	Medium Shell Magnal 11 Pin	11N 11N	Green White	Medium Medium	1000	1500	... .	550	1000 500	163-291 245-437	. . ....	16.5-49.5 25-75	64-96 96-144	56-84 84-126
3JP1	6.3	0.6	3	10	Medium Shell Diheptal 12 Pin	14J 14J	Green	Medium	1000	2000	4000	500	500 2000 500 2000	430 575 430 575	1500 2000 3000 4000	22.5-67.5 30-90 22.5-67.5 30-90	120 160 150 200	89 119 111 148
3KP1 3KP4	6.3	0.6	3	11-1/2	Medium Shell Magnal 11 Pin	11M 11M	Green White	Medium Medium	1000	2500		500	1000 2000	160-300 320-600	....	0-45 0-90	50-68 100-136	38-52 76-104
3MP1	6.3	0.6	3	8	Small Shell Duodecal 12 Pin	12F	Green	Medium	1000	2500	..	500	1000 2000	200-350 400-700	....	0-63 0-126	140-190 280-380	130-180 260-360
3QP1	6.3	0.3	2-3/4	6-1/8	European 9 Pin	9D	Green	Medium	700	1500	...	550	800 1200	200-320 240-480	....	21-50 31-74	143-193 214-290	89-121 133-181
3RP1 3RP1A	6.3	0.6	3	9-1/8	Small Shell Duodecal 12 Pin	12E	Green	Medium	1000	2500	. ..	500	1000 2000	165-310 330-620	....	67.5 13.5	85 61	172 122
3SP1 3SP4	6.3	0.6	3x1-1/2	9-1/8	Small Shell Duodecal 12 Pin	12E	Green White	Medium Medium	1100	2750	....	...	1000 2000	165-310 330-620	....	28.5-67.5 58-135	73-99 146-198	52-70 104-140
5AP1	6.3	0.6	5-1/4	13	Large Wafer Magnal 11 Pin, Sleeve	11A	Green	Medium	1200	2000	....	500	1500 2000	430 575	....	31-57 40-74	93	98
5AP4	6.3	0.6	5-1/4	13	Large Wafer Magnal 11 Pin, Sleeve	11A	White	Medium	1200	2000	....	500	1500 2000	430 575	.. . .	17.6-57 22.8-74	93	98
5BP1 5BP4	6.3	0.6	5-1/4	16-3/4	Large Wafer Magnal 11 Pin	11A 11A	Green White	Medium Medium	1000	2000	....	500	1500 2000	310 425	....	20-60	63 84	57 76
5BP1A	6.3	0.6	5-1/4	16-3/4	Medium Shell Magnal 11 Pin	11N	Green	Medium	1000	2000	....	500	1500 2000	337-450	....	15-45 20-60	63 84	57 76
5BP7A	6.3	0.6	5-1/4	16-3/4	Medium Shell Magnal 11 Pin	11N	See chap.7 for Characteristics of P7 Screen		1000	2000	....	500	1500 2000	235-420 315-560	....	15-45 20-60	52-74 70-98	47-67 63-89

*Cut-off voltage. Supply should be adjustable from 0 to value shown. **Bogey value for focus. Voltage should be adjustable about value shown.

Commonly used Phosphors only listed.

Courtesy Sylvania Elec. Prod. Inc.

ELECTROSTATIC TYPES—CATHODE RAY TUBES

Type	Heater		Nominal Dimensions		Base	RMA Basing	Screen		Maximum Design Center Ratings				Typical Operating Conditions					
	Volts	Amperes	Diameter Inches	Length Inches			Fluorescence	Persistence	Anode #1 Volts	Anode #2 Volts	Anode #3 Volts	Anode #2 to Deflection Plate Peak Volts	Anode #2 Volts	Anode #1 Avg Volts**	Anode #3 Volts	Grid Range Volts*	Deflection Avg. Volts DC/Inch D 1-2	Deflection Avg. Volts DC/Inch D 3-4
5CP1 5CP4	6.3	0 6	5-1/4	16-3/4	Medium Shell Diheptal 12 Pin	14B 14B	Green White	Medium Medium	1000	2000	4000	500	2000 1500 2000	575 430 575	2000 3000 4000	30-90 22 5-67 5 30-90	73 69 92	64 56 74
5CP1A	6.3	0 6	5-1/4	16-3/4	Medium Shell Diheptal 12 Pin	14J 14J	Green	Medium	1000	2000	4000	500	2000 1500 2000	575 430 575	2000 3000 4000	30-90 22 5-67 5 30-90	73 69 92	64 56 74
5GP1	6.3	0.6	5-1/4	16-3/4	Large Wafer Magnal 11 Pin, Sleeve	11A	Green	Medium	1000	2000		500	2000	425		24-56	36	72
5HP1 5HP4	6.3	0.6	5-1/4	16-3/4	Large Wafer Magnal 11 Pin, Sleeve	11A 11A	Green White	Medium Medium	1000	2000		500	1500 2000	310 425		15-45 20-60	63 5 84 8	57 8 77 0
5HP1A	6.3	0.6	5-1/4	16-3/4	Large Wafer Magnal 11 Pin, Micanol	11N	Green	Medium	1000	2000		500	1500 2000	337 450		15-45 20-60	63 84	57 76
5JP1 5JP4	6.3	0.6	5-5/16	16-3/4	Medium Magnal 11 Pin	11E 11E	Green White	Medium Medium	1000	2000	4000	500	1000 2000	260 520	2000 4000	22 2-51 8 45-105	96	96
5JP1A 5JP4A	6.3	0.6	5-5/16	16-3/4	Medium Magnal 11 Pin	11S 11S	Green White	Medium Medium	1000	2000	4000	500	1500 2000	250-472 333 630	3000 4000	34-79 45-105	58-86 77-115	58-86 77-115
5LP1 5LP4	6.3	0.6	5-5/16	16-3/4	Medium Magnal 11 Pin, Sleeve	11F 11F	Green White	Medium Medium	1000	2000	4000	500	1000 1500 2000	250 375 500	2000 3000 4000	15-45 22 5-67 5 30-90	52 77 103	45 68 90
5LP1A 5LP4A	6.3	0.6	5-5/16	16-3/4	Medium Magnal 11 Pin, Sleeve	11T 11T	Green White	Medium Medium	1000	2000	4000	550	1500 2000	282 475 376-633	3000 4000	22 5-67 5 30-90	62-93 83-124	54-81 72-108
5MP1 5MP4	2.5	2.1	5-5/16	15-7/8	Large 7 Pin	7AN 7AN	Green White	Medium Medium	1000	1500		600	1000 1500	250 375		16 5-49 5 15-45	66	60
5NP1 5NP4	6.3	0.6	5-5/16	16-3/4	Large Wafer Magnal 11 Pin, Sleeve	11A 11A	Green White	Medium Medium	1000	2000		500	1500 2000	337 450		15-45 20-60	84	76
5RP1 5RP4	6.3	0.6	5-1/4	16-3/4	Medium Shell Diheptal 12 Pin	14F 14F	Green White	Medium Medium	15550	3500	25500	1200	2000 2000	518 528	10000 20000	30-90 30-90	30-45 36-54	30-45 36-54
5RP1A 5RP4A	6.3	0.6	5-1/4	16-3/4	Medium Shell Diheptal 12 Pin	14F	Green White	Medium Medium	15550	3500	25500	1200	2000 2000	518 528	10000 20000	30-90 30-90	30-45 36-54	30-45 36-54
5SP1 5SP4	6.3	0.6	5-1/4	18-1/2	Medium Shell Diheptal 12 Pin	14K 14K	Green White	Medium Medium	1000	2000	4000	500	1500 1500 2000	431 431 575	1500 3000 4000	22 5-67 5 22 5-67 5 30-90	55 69 92	48 59 79
5UP1	6.3	0.6	5-1/4	14-3/4	Small Shell Duodecal 12 Pin	12F	Green	Medium	1000	2500		500	1000 2000	170-320 340-640		22 5-67 5 30-90	28-38 5 56-77	28-31 46-62
5VP7	6.3	0.6	5-1/4	16-3/4	Medium Shell Magnal 11 Pin	11N	See chap.7 for Characteristics of Phosphor No 7		1000	2500		500	1500 2000	236 422 315-562		15-45 20-60	52-74 70-98	47-67 63-89
5XP1	6.3	0.6	5-1/4	17-5/8	Medium Shell Diheptal 12 Pin	14F	Green	Medium	1550	3500	25500	1200	2000 2000 2000	362-695 362-695 362-695	4000 10000 20000	30-90 30-90 30-90	72-108 102-695 362-695	24-36 34-52 46-68
7EP4	6.3	0.6	7	15-1/2	Medium Shell Magnal 11 Pin	11N	White	Medium	1500	3300		700	2500	650		36-84	110	95
7GP4	6.3	0.6	7	14-1/2	Medium Shell Diheptal 12 Pin	14G	White	Medium	1500	4000		500	3000	810-1200		36-84	93-123	75-102
7JP1 7JP4	6.3	0.6	7	14-1/2	Medium Shell Diheptal 12 Pin	14G	Green White	Medium Medium	2800	6000		750	6000	1620-2400		72-168	186-246	150-204
[illegible]	6.3	0.6	8-3/4	16-1/2	Medium Shell Diheptal 12 Pin	14G	White	Medium	3100	6600		750	6000	2000		72-168	146-198	124-198
[illegible]	2.5	2.1	9	21	Medium 6 Pin	6BN	Green	Medium	1500	5500		1500	5000	1150		45-135	190	175

* Cut-off voltage. Supply should be adjustable from 0 to value shown

** Bogey value for focus Voltage should be adjustable about value shown

Courtesy Sylvania Elec Prod. Inc

ELECTROSTATIC TYPES—CATHODE RAY TUBES

Type	Heater		Nominal Dimensions		Base	RMA Basing	Screen		Maximum Design Center Ratings				Typical Operating Conditions					
	Volts	Amperes	Diameter Inches	Length Inches			Fluorescence	Persistence	Anode #1 Volts	Anode #2 Volts	Anode #3 Volts	Anode #2 to Deflection Plate Peak Volts	Anode #2 Volts	Anode #1 Avg. Volts**	Anode #3 Volts	Grid Range Volts*	Deflection Avg. Volts DC/Inch D 1-2	Deflection Avg. Volts DC/Inch D 3-4
10SP4	6.3	0.6	10	18-1/2	Medium Shell Diheptal 12 Pin	14G	White	Medium	2000	5000	....	500	4500 5000	1130-1660 1250-1850	....	54-126 60-140	112-149 125-165	90-127 100-135
10MP4	6.3	0.6	10	19-1/4	Medium Shell Diheptal 12 Pin	14G	White	Medium	2000	5000	....	600	4000 5000	960-1440 1200-1800	....	48-112 60-140	88-120 110-150	68-92 85-115
12SP7	6.3	0.6	12	24	Medium Shell Diheptal 12 Pin	14E	See chap.7 for Characteristics of Phosphor No. 7		2000	4000	8000	1000	2000 4000 3000 4000	625 1250 937 1250	4000 4000 6000 8000	30-90 30-90 30-90 30-90	55 83 110	63 94 125
12GP7	6.3	0.6	12	22	Medium Shell Diheptal 12 Pin	14B	See chap.7 for Characteristics of Phosphor No. 7		2000	4000	6000	1000	3000 3000 4000 4000	857 857 1143 1143	3000 6000 4000 6000	49-147 49-147 65-195 65-195	73 89 97 108	68 83 91 101
12MP1	6.3	0.6	12	23-1/2	Medium Magnal 11 Pin, Sleeve	11J	Green	Medium	1500	5500	....	1000	5000	1150 +25% -30%	....	45-135	19	25
14GP1 14GP4	2.5	2.1	13-3/8	24-1/4	12 Pin Peripheral Contact	12A 12A	Green White	Medium Medium	1800	4000	8000	....	2000 4000	500 1000	4000 8000	20-60 40-120	65 130	65 130
20AP1 20AP4	2.5 2.5	2.1 2.1	20 20	27-7/8 27-7/8	12 Pin Peripheral Contact	12A 12A	Green White	Medium Medium	1800 1800	4000 4000	8000 8000	....	2000 4000 2000 4000	500 1000 500 1000	4000 8000 4000 8000	20-60 40-120 20-60 40-120	55 110 65 130	55 110 65 130
902	6.3	0.6	2	7-1/2	Medium Shell Octal 8 Pin	8CD	Green	Medium	300	600	...	347	400 600	100 150	....	20-60 30-90	93 139	78 117
902-A	6.3	0.6	2	7-7/16	Medium Shell Octal 8 Pin	8CD	Green	Medium	300	600	....	347	400 600	100 150	....	20-60 30-90	93 139	78 117
905 907 909	2.5	2.1	5-1/4	16-1/2	Long Shell Medium 5 Pin Micanol	5BP 5BP 5BP	Green Blue Bluish-White	Medium Very Short Long	600	2000	....	1000	1500 2000	338 450	....	13-39 17.5-52.5	86 115	73 97
905-A	2.5	2.1	5-1/4	16-1/2	Long Shell Medium 5 Pin Micanol	5BR	Green	Medium	600	2000	....	1000	1500 2000	338 450	....	13-39 17.5-52.5	86 115	73 97
908 910	2.5	2.1	3	11-1/2	Medium 7 Pin	7AN 7AN	Bluish Bluish-White	Very Short Long	1000	1500	....	600	600 800 1000 1200 1500	170 230 285 345 475	....	13-46 30-70	46.3 62 77 94 115.2	44 57.8 72.5 88 110
908-A	2.5	2.1	3	11-1/2	Medium 7 Pin	7CE	Blue	Very Short	1000	1500	....	500	1000 1500	287 430	....	16.5-49.5 25-75	76 114	73 109
912	2.5	2.1	5-1/4	16-1/2	Medium 5 Pin Micanol	912	Green	Medium	4500	1500	....	7000	5000 10,000 15,000	1000 2000 3000	....	27-81 31-93 35-105	306 620 910	248 498 746
913	6.3	0.6	1-5/8	4-3/4	Small Wafer Octal 8 Pin	913	Green	Medium	200	500	....	250	250 500	50 100	....	10-30 32-98	169 363	121 254
914	2.5	2.1	9-1/4	21-1/2	Medium 6 Pin	6BF	Green	Medium	1900	7000	....	3000	1500 2500 5000 7000	300 515 1030 1450	....	25-75 25-75 25-75 25-75	75 124.5 248 348	58.7 97.8 195 274
914A	2.5	2.1	9-1/4	20-1/16	Medium 6 Pin	914A	Green	Medium	1900	7000	....	3000	1500 2500 5000 7000	320 550 1100 1550	....	25-75 25-75 25-75 25-75	69.5 115 231 323	54.6 91 182 254
VCR 139A	4.0	1.1	2-3/4	7-7/8	European	VCR 139A	Green	Medium	1000	1000	....		800	120-150	....	7-16	104	140

* Cut-off voltage. Supply should be adjustable from 0 to value shown.

** Bogey value for focus. Voltage should be adjustable about value shown.

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MAGNETIC TYPE CATHODE RAY TUBES

Type No.	Heater		Bulb						Ion Trap Required	Base	RMA Basing	μμf Filter Capacitance Provided by Bulb Coating	Deflection and Focusing Method	Maximum Design Center Ratings		Typical Operation			Type No
	Volts	Amperes	Nominal Face Dimensions in Inches	Length in Inches	Construction	Terminal	Face Plate Color	Deflection Angle in Degrees (Note 1)						Anode Volts	Accelerator Grid Volts	Anode Volts	Accelerator Grid Volts	Control Grid Negative Volts	
3WP7	6.3	0.6	3 Diam.	9-13/16	Glass	Snap	Clear	55	None	Medium Shell Octal 8 Pin	5AN	None	Magnetic	5000	200	4000 5000	150 150	15-45 15-45	3WP7
3NP4	6.3	0.6	2-9/16 Diam.	10	Glass	Recessed Small Ball	Clear	42	None	Special 5 Pin	3NP4	275 Min, 375 Max	Magnetic	25000	.	24000	..	60	3NP4
5FP4A	6.3	0.6	5 Diam	11-1/8	Glass	Recessed Small Ball	Clear	53	None	Medium Shell Octal 8 Pin	8BX	None	Magnetic	8000	300	6000	250	45	5FP4A
5FP7A	6.3	0.6	5 Diam.	11-1/2	Glass	Recessed Small Ball	Clear	53	None	Medium Shell Octal 8 Pin	8BX	None	Magnetic	8000	700	4000 7000	250 250	25-70 25-70	5FP7A
5FP7	6.3	0.6	5 Diam.	11-1/8	Glass	Snap	Clear	55	None	Small Wafer Octal 8 Pin with Sleeve	5AN	None	Magnetic	7000	300	4000 7000	250 250	25-75 25-75	5FP7
5FP14	6.3	0.6	5 Diam.	11-1/8	Glass	Snap	Clear	55	None	Small Wafer Octal 8 Pin with Sleeve	5AN	None	Magnetic	7000	700	4000 7000	250 250	25-75 25-75	5FP14
5TP4	6.3	0.6	5 Diam.	11-3/4	Glass	Recessed Small Cavity	Clear	50	None	Medium Shell Diheptal 12 Pin	12C1	100 Min, 500 Max	Note 2	27000	350	27000	200	70	5TP4
5WP15	6.3	0.6	5 Diam.	11-7/16	Glass	Recessed Small Cavity	Clear	50	None	Small Shell Duodecal 7 Pin	12C1	100 Min, 500 Max	Note 2	27000	350	20000	200	70	5WP15
7AP4	2.5	2.1	7-1/8 Diam.	13-1/2	Glass	None	Clear	55	None	Medium 5 Pin	5AJ	None	Note 2	35000	No Grid	35000	No Grid	67 5	7AP4
7BP1	6.3	0.6	7 Diam.	13-1/4	Glass	Snap	Clear	55	None	Octal 8 Pin with Sleeve	5AN	None	Magnetic	7000	675	4000 7000	250 250	50 50	7BP1
7BP7	6.3	0.6	7 Diam.	13-1/4	Glass	Snap	Clear	55	None	Octal 8 Pin with Sleeve	5AN	None	Magnetic	7000	300	4000 7000	250 250	50 50	7BP7
7BP7A	6.3	0.6	7 Diam.	13-1/4	Glass	Recessed Small Ball	Clear	53	None	Medium Shell Octal 8 Pin	8BX	None	Magnetic	8000	700	4000 7000	250 250	25-70 25-70	7BP7A
7CP1	6.3	0.6	7 Diam	13-7/16	Glass	Snap	Clear	57	None	Medium Shell Octal 8 Pin	6A7	None	Note 2	8000	300	4000 7000	250 250	45 45	7CP1
7CP4	6.3	0.6	7 Diam	13-7/16	Glass	Recessed Small Ball	Clear	57	None	Medium Shell Octal 8 Pin	6AZ	None	Note 2	8000	300	6000	250	45	7CP4
7DP4	6.3	0.6	7-3/16 Diam.	14-1/16	Glass	Recessed Small Cavity	Clear	50	Double	Small Shell Duodecal 7 Pin	12C2	400 Min, 1500 Max.	Note 2	8000	410	6000	250	45	7DP4
7EP4	6.3	0.6	7-3/16 Diam.	13	Glass	Recessed Small Ball	Clear	50	None	Small Shell Duodecal 7 Pin	12D2	500 Max	Magnetic	8000	410	6000	250	33-77	7EP4
8AP4	6.3	0.6	8-1/2 Diam.	14-1/4	Metal	Cone Lip	Clear	54	Single	Small Shell Duodecal 7 Pin	12H	None	Magnetic	10000	No Grid	9000	No Grid	27-63	8AP4
8AP4A	6.3	0.6	8-1/2 Diam.	14-1/4	Metal	Cone Lip	Gray	54	Single	Small Shell Duodecal 5 Pin	12H	None	Magnetic	9000	No Grid	7000	No Grid	27-63	8AP4A
9AP4	2.5	2.1	9-1/8 Diam.	21	Glass	Cap	Clear	40	None	Medium 6 Pin	6AL	None	Note 2	7000	250	6000 7000	250 250	25 25	9AP4
9CP4	2.5	2.1	9 Diam.	15-7/8	Glass	Cap	Clear		None	6 Pin Base	4AF	None	Magnetic	7000	No Grid	6000 7000	No Grid	90 100	9CP4
9GP7	6.3	0.6	9 Diam.	17	Glass	Cap.	Clear	55	None	Octal 8 Pin with Sleeve	5AN	None	Magnetic	7000	300	4000 7000	250 250	45 45	9GP7
9JP1	2.5	2.1	9 Diam.	15-11/16	Glass	Snap	Clear	55	None	Small Wafer Octal 8 Pin with Sleeve	8BR	None	Note 3	5000	No Grid	2500 5000	No Grid	45 90	9JP1
9LP7	6.3	0.6	9 Diam.	14-31/32	Glass	Cap	Clear	55	None	Octal 8 Pin with Sleeve	5AN	None	Magnetic	7000	300	4000 4000 7000 7000	250 125 250 125	55-105 30-50 60-100 30-50	9LP7
9MP7	6.3	0.6	9 Diam.	17-1/2	Glass	Cap	Clear	55	None	Octal 8 Pin with Sleeve	5AN	None	Magnetic	7000	300	4000 6000	250	25-75	9MP7
10BP4	6.3	0.6	10-1/2 Diam.	17-5/8	Glass	Recessed Small Cavity	Clear	50	Double	Small Shell Duodecal 7 Pin	12D2	500 Min, 2500 Max	Magnetic	10000	410	9000	250	20-60	10BP4
10BP4A							Gray												10BP4A
10CP4	6.3	0.6	10-1/2 Diam.	16-5/8	Glass	Recessed Small Ball	Clear	50	None	Small Shell Duodecal 7 Pin	12D2	500 Max	Magnetic	11000	410	8000	250	30-66	10CP4

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MAGNETIC TYPE CATHODE RAY TUBES

Type No.	Heater		Bulb						Ion Trap Required	Base	RMA Basing	µµf Filter Capacitance Provided by Bulb Coating	Deflection and Focusing Method	Maximum Design Center Ratings		Typical Operation			Type No.
	Volts	Amperes	Nominal Face Dimensions in Inches	Length in Inches	Con-struction	Terminal	Face Plate Color	Deflection Angle in Degrees (Note 1)						Anode Volts	Acceler-ator Grid Volts	Anode Volts	Acceler-ator Grid Volts	Control Grid Negative Volts	
10BP4	6.3	0.6	10-1/2 Diam.	17-5/8	Glass	Recessed Small Cavity	Clear	50	None	Small Shell Duodecal 7 Pin	12C3	None	Note 2	10000	410	9000	250	36-84	10BP4
10EP4	6.3	0.6	10-1/2 Diam.	17-5/8	Glass	Snap	Clear	50	Double	Small Shell Duodecal 7 Pin	12D2		Magnetic	11000	330	8000	250	20-65	10EP4
10FP4	6.3	0.6	10-1/2 Diam.	17-5/8	Glass	Recessed Small Cavity	Clear	50	None	Small Shell Duodecal 7 Pin	12C1	500 Min 2500 Max	Magnetic	10000	410	9000	250	27-63	10FP4
10KP7	6.3	0.6	10-1/2 Diam.	17-5/8	Glass	Recessed Small Cavity	Clear	50	None	Small Shell Duodecal 7 Pin	12D1	None	Magnetic	10000	700	7000 9000	250 250	27-63 27-63	10KP7
10MP4	6.3	0.6	10-1/2 Diam.	17	Glass	Recessed Small Cavity	Clear	52	Double	Small Shell Duodecal 5 Pin	12G	500 Min, 2500 Max	Magnetic	10000	No Grid	9000	No Grid	27-63	10MP4
10MP4A							Gray												10MP4A
12AP4	2.5	2.1	12-1/16 Diam.	25-3/8	Glass	Cap	Clear	40	None	Medium 6 Pin	6AL	None	Note 2	7000	250	6000 7000	250 250	75 75	12AP4
12CP4	2.5	2.1	12-1/16 Diam.	18-5/8	Glass	Cap	Clear		None	6 Pin Base	4AF	None	Magnetic	7000	No Grid	6000 7000	No Grid	90 110	12CP4
12DP7	6.3	0.6	12 Diam.	20-3/4	Glass	Medium Cap	Clear	55	None	Small Wafer Octal 8 Pin with Sleeve	5AN	None	Magnetic	7000	300	4000 7000	250 250	25-75 25-75	12DP7
12DP7A	6.3	0.6	12 Diam.	19-5/8	Glass	Medium Cap	Clear	50	None	Medium Shell Octal 8 Pin	8BX	None	Magnetic	10000	700	4000 7000	250 250	25-70 25-70	12DP7A
12JP4	6.3	0.6	12 Diam.	17-1/2	Glass	Snap	Clear	50	None	Small Shell Duodecal 7 Pin	12D1	None	Magnetic	12000	410	10000	250	27-63	12JP4
12KP4	6.3	0.6	12-7/16 Diam.	17-5/8	Glass	Recessed Small Cavity	Clear	54	None	Small Shell Duodecal 7 Pin	12D2	500 Min, 2500 Max	Magnetic	12000	410	10000	250	27-63	12KP4
12KP4A	6.3	0.6	12-7/16 Diam	17-5/8	Glass	Recessed Small Cavity	Gray	54	None	Small Shell Duodecal 5 Pin	12D2	500 Min, 2500 Max	Magnetic	12000	410	11000	250	27-63	12KP4A
12LP4	6.3	0.6	12-7/16 Diam.	18-3/4	Glass	Recessed Small Cavity	Clear	54	Double	Small Shell Duodecal 5 Pin	12D2	750 Min, 3000 Max	Magnetic	12000	410	11000	250	27-63	12LP4
12LP4A							Gray												12LP4A
12QP4	6.3	0.6	12-7/16 Diam.	17-1/2	Glass	Recessed Small Ball Cap	Clear	55	Single	Small Shell Duodecal 7 Pin	12D1	None	Magnetic	12000	410	10000	250	27-63	12QP4
12QP4A							Gray												12QP4A
12RP4	6.3	0.6	12 Diam	17-1/2	Glass	Recessed Small Ball Cap	Clear	56	Single	Small Shell Duodecal 7 Pin	12D2		Magnetic	12000	410	10000	250	27-63	12RP4
12SP7	6.3	0.6	12-7/16 Diam	18-3/4	Glass	Recessed Small Cavity	Clear	55	None	Small Shell Duodecal 7 Pin	12D1	None	Magnetic	10000	410	9000	250	27-63	12SP7
12TP4	6.3	0.6	12-7/16 Diam	18-3/4	Glass	Recessed Small Cavity	Clear	54	Double	Small Shell Duodecal 7 Pin	12D1	None	Magnetic	12000	410	11000	250	27-63	12TP4
12UP4	6.3	0.6	12-7/16 Diam	18-5/8	Metal	Cone Lip	Clear	54	Double	Small Shell Duodecal 7 Pin	12D3	None	Magnetic	12000	410	11000	250	27-63	12UP4
12UP4A							Gray												12UP4A
12UP4B							Gray												12UP4B
12VP4	6.3	0.6	12-3/8 Diam.	18	Glass	Recessed Small Cavity	Clear	55	Double	Small Shell Duodecal 5 Pin	12G	750 Min, 3000 Max	Magnetic	12000	No Grid	11000	No Grid	33-77	12VP4
12VP4A							Gray												12VP4A
14BP4	6.3	0.6	12-1/2 x 9-11/16	16-13/16	Glass	Recessed Small Cavity	Gray	65	Double	Small Shell Duodecal 5 Pin	12D2	500 Min, 2000 Max	Magnetic	12000	410	11000	250	27-63	14BP4
14CP4	6.3	0.6	12-1/2 x 9-11/16	16-3/4	Glass	Recessed Small Cavity	Gray	65	Double	Small Shell Duodecal 5 Pin	12D2	1500	Magnetic	14000	410	12000	300	33-77	14CP4
14DP4	6.3	0.6	12-1/2 x 9-11/16	16-3/4	Glass	Recessed Small Cavity	Gray	65	Double	Small Shell Duodecal 5 Pin	12D1	None	Magnetic	14000	410	11000	250	27-63	14DP4
15AP4	6.3	0.6	15-1/2 Diam.	20-1/2	Glass	Recessed Small Ball	Clear	52	None	Small Shell Duodecal 7 Pin	12G	None	Magnetic	15000	410	12000	250	27-63	15AP4
15CP4	6.3	0.6	15-1/2 Diam.	21-1/2	Glass	Recessed Small Cavity	Clear	57	Double	Small Shell Duodecal 7 Pin	12D1	None	Magnetic	15000	410	9000 15000	250 +	45	15CP4

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MAGNETIC TYPE CATHODE RAY TUBES

Type No.	Heater		Bulb						Ion Trap Required	Base	RMA Basing	µµf Filter Capacitance Provided by Bulb Coating	Deflection and Focusing Method	Maximum Design Center Ratings		Typical Operation			Type No
	Volts	Amperes	Nominal Face Dimensions in Inches	Length in Inches	Construction	Terminal	Face Plate Color	Deflection Angle in Degrees (Note 1)						Anode Volts	Accelerator Grid Volts	Anode Volts	Accelerator Grid Volts	Control Grid Negative Volts	
15DP4	6.3	0.6	15-1/2 Diam	20-1/2	Glass	Recessed Small Ball Cap	Clear	57	Double	Small Shell Duodecal 5 Pin	12D1	None	Magnetic	15000	410	13000	250	27-63	15DP4
16AP4	6.3	0.6	15-7/8 Diam	22-5/16	Metal	Cone Lip	Clear	53	Double	Small Shell Duodecal 5 Pin	12D3	None	Magnetic	14000	410	9000 12000	300 300	33-77 33-77	16AP4
16AP4A							Gray												16AP4A
16CP4	6.3	0.6	15-7/8 Diam	21-1/2	Glass	Recessed Small Cavity	Clear	52	Double	Small Shell Duodecal 7 Pin	12D1	None	Magnetic	15000	410	12000	250	27-63	16CP4
16DP4	6.3	0.6	15-7/8 Diam.	20-3/4	Glass	Recessed Small Cavity	Clear	60	Double	Small Shell Duodecal 7 Pin	12D1	None	Magnetic	15000	410	9000 12000	250	45	16DP4
16DP4A							Gray												16DP4A
16EP4	6.3	0.6	15-7/8 Diam.	19-5/8	Metal	Cone Lip	Clear	60	Double	Small Shell Duodecal 5 Pin	12D3	None	Magnetic	14000	410	12000	300	33-77	16EP4
16EP4A							Gray												16EP4A
16FP4	6.3	0.6	16-1/8 Diam.	20-1/4	Glass	Recessed Small Ball Cap	Clear	62	Single	Small Shell Duodecal 7 Pin	12D1	None	Magnetic	16000	410	13000	250	27-63	16FP4
16GP4	6.3	0.6	15-7/8 Diam.	17-11/16	Metal	Cone Lip	Clear	70	Single	Small Shell Duodecal 5 Pin	12D3	None	Magnetic	14000	410	12000	300	33-77	16GP4
16HP4	6.3	0.6	15-7/8 Diam.	21-1/4	Glass	Recessed Small Cavity	Clear	60	Double	Small Shell Duodecal 5 Pin	12D2	1500 Min., 3500 Max.	Magnetic	14000	410	12000	300	33-77	16HP4
16HP4A							Gray												16HP4A
16JP4	6.3	0.6	16-1/8 Diam	20-3/4	Glass	Recessed Small Cavity	Clear	60	Double	Small Shell Duodecal 5 Pin	12D2	750 Min., 2000 Max.	Magnetic	14000	410	11000	250	27-63	16JP4
16JP4A							Gray												16JP4A
16KP4	6.3	0.6	14-3/4 x 11-1/2	18-3/4	Glass	Recessed Small Cavity	Clear	65	Single	Small Shell Duodecal 5 Pin	12D2	1500	Magnetic	16000	410	14000	300	33-77	16KP4
16LP4	6.3	0.6	15-7/8 Diam	22-1/4	Glass	Recessed Small Cavity	Clear	52	Double	Small Shell Duodecal 5 Pin	12D2	1500 Min., 3500 Max.	Magnetic	14000	410	12000	300	33-77	16LP4
16LP4A							Gray												16LP4A
16MP4	6.3	0.6	16-1/8 Diam.	21-3/4	Glass	Recessed Small Cavity	Clear	60	Double	Small Shell Duodecal 5 Pin	12D2	1500 Min., 3500 Max	Magnetic	14000	410	12000	300	33-77	16MP4
16MP4A							Gray												16MP4A
16QP4	6.3	0.6	14-3/4 x 11-17/32	19 1/16	Glass	Recessed Small Cavity	Gray	65	Double	Small Shell Duodecal 7 Pin	12D1	None	Magnetic	16000	410	8000 14000	250 250	27-63	16QP4
16RP4	6.3	0.6	14-3/4 x 11-1/2	18-3/4	Glass	Recessed Small Cavity	Gray	65	Single	Small Shell Duodecal 5 Pin	12D2	1500	Magnetic	16000	410	12000	300	33-77	16RP4
16SP4	6.3	0.6	15-7/8 Diam	17-5/16	Glass	Recessed Small Cavity	Clear	70	Double	Small Shell Duodecal 5 Pin	12D2	1500 Min., 3500 Max.	Magnetic	14000	410	12000	300	33-77	16SP4
16SP4A							Gray												16SP4A
16TP4	6.3	0.6	16-1/8 Diam.	18-1/8	Glass	Recessed Small Cavity	Gray	70	Single	Small Shell Duodecal 5 Pin	12D2	1500	Magnetic	14000	410	12000	300	33-77	16TP4
16UP4	6.3	0.6	14-3/4 x 11-1/2	18-1/8	Glass	Recessed Small Cavity	Gray	65	Single	Small Shell Duodecal 5 Pin	12D1	None	Magnetic	15000	410	12000	300	27-63	16UP4
16VP4	6.3	0.6	15-7/8 Diam.	17-3/16	Glass	Recessed Small Cavity	Gray	70	Single	Small Shell Duodecal 5 Pin	12D1	None	Magnetic	15000	410	12000	250	27-63	16VP4
16WP4	6.3	0.6	15-7/8 Diam.	17-3/4	Glass	Recessed Small Cavity	Gray	70	Double	Small Shell Duodecal 5 Pin	12D1	None	Magnetic	15000	410	12000	250	27-63	16WP4
16XP4	6.3	0.6	14-3/4 x 11-17/32	18-3/4	Glass	Recessed Small Cavity	Gray	65	Double	Small Shell Duodecal 5 Pin	12D1	None	Magnetic	15000	410	12000	250	27-63	16XP4
16YP4	6.3	0.6	15-7/8 Diam.	17-5/16	Glass	Recessed Small Cavity	Gray	70	Single	Small Shell Duodecal 5 Pin	12D2	750 Min., 2000 Max.	Magnetic	14000	410	12000	300	33-77	16YP4
19AP4	6.3	0.6	18-5/8 Diam.	21-1/2	Metal	Cone Lip	Clear	66	Single	Small Shell Duodecal 7 Pin	12D3	None	Magnetic	19000	410	13000	250	27-63	19AP4
19AP4A							Gray												19AP4A

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MAGNETIC TYPE CATHODE RAY TUBES

Type No.	Heater		Bulb						Ion Trap Required	Base	RMA Basing	μμf Filter Capacitance Provided by Bulb Coating	Deflection and Focusing Method	Maximum Design Center Ratings		Typical Operation			Type No.
	Volts	Amperes	Nominal Face Dimensions in Inches	Length in Inches	Construction	Terminal	Face Plate Color	Deflection Angle in Degrees (Note 1)						Anode Volts	Accelerator Grid Volts	Anode Volts	Accelerator Grid Volts	Control Grid Negative Volts	
19DP4	6.3	0.6	18-7/8 Diam.	21-1/2	Glass	Recessed Small Cavity	Clear	66	Double	Small Shell Duodecal 5 Pin	12D2	1000 Min, 3000 Max.	Magnetic	19000	410	13000	250	26-63	19DP4
19FP4	6.3	0.6	18-7/8 Diam.	22	Glass	Recessed Small Cavity	Gray	66	Double	Small Shell Duodecal 5 Pin	12D1	None	Magnetic	19000	410	13000	250	27-63	19FP4
19GP4	6.3	0.6	18-7/8 Diam.	21-1/4	Glass	Recessed Small Cavity	Gray	66	Single	Small Shell Duodecal 5 Pin	12D1	None	Magnetic	19000	410	13000	250	27-63	19GP4
20BP4	6.3	0.6	20 Diam.	28-3/4	Glass	Medium Cap	Clear	54	None	Small Shell Duodecal 7 Pin	12D1	None	Magnetic	16500	750	10000 15000	250 250	25-70 25-70	20BP4
22AP4	6.3	0.6	21-11/16 Diam.	22-7/8	Metal	(Cone Lip)	Clear	70	Single	Small Shell Duodecal 5 Pin	12D3	None	Magnetic	19000	410	14000	300	33-77	22AP4
22AP4A							Gray												22AP4A
904	2.5	2.1	5-1/16 Diam.	16-1/4	Glass	Cap	Clear		None	Medium 6 Pin	6AL	None	Note 4	4600	250	1000 3000 4600	100 100 250	34 35 39	904
5WP11	6.3	0.6	5 Diam.	11-7/16	Glass	Recessed Small Cavity	Clear	50	None	Small Shell Duodecal 7 Pin	12C2	100 Min, 500 Max	Note 2	27000	350	27000	200	42-98	5WP11
7MP7	6.3	0.6	7-3/16 Diam.	12-1/2	Glass	Recessed Small Cavity	Clear	50	None	Small Shell Duodecal 5 Pin	12D1	None	Magnetic	8000	700	4000 7000	250 250	27-63 27-63	7MP7
19EP4	6.3	0.6	17 x 13-3/32	21-1/8	Glass	Recessed Small Cavity	Gray	65	Single	Small Shell Duodecal 5 Pin	12D2	1000 Min, 2500 Max.	Magnetic	19000	410	13000	250	26-63	19EP4
16ZP4	6.3	0.6	15-7/8 Diam.	22-1/4	Glass	Recessed Small Cavity	Gray	52	Single	Small Shell Duodecal 5 Pin	12D2	750 Min, 2000 Max	Magnetic	16000	410	12000	300	33-77	16ZP4
16WP4A	6.3	0.6	15-7/8 Diam.	17-3/4	Glass	Recessed Small Cavity	Gray	70	Single	Small Shell Duodecal 5 Pin	12D2	750 Min, 2000 Max	Magnetic	16000	410	12000	250	27-63	16WP4A
17AP4	6.3	0.6	15-3/8 x 12-1/4	18-5/8	Glass	Recessed Small Cavity	Gray	65	Single	Small Shell Duodecal 5 Pin	12D2	750 Min, 2000 Max	Magnetic	16000	410	12000	300	33-77	17AP4
17BP4	6.3	0.6	15-25/64 x 12-9/64	19-5/8	Glass	Recessed Small Cavity	Clear	65	Single	Small Shell Duodecal 5 Pin	12D2	750 Min, 2000 Max.	Magnetic	16000	410	12000	300	33-77	17BP4
17BP4A							Gray												17BP4A
10FP4A	6.3	0.6	10-1/2 Diam.	17-5/8	Glass	Recessed Small Cavity	Gray	54	None	Small Shell Duodecal 5 Pin	12D2	500 Min, 2500 Max.	Magnetic	12000	410	11000	250	27-63	10FP4A

Note 1· Horizontal Deflection Angles are given for Rectangular Tubes

Note 2· Magnetic Deflection, Electrostatic Focusing.

Note 3 Electrostatic and Magnetic Deflection, Magnetic Focusing.

Note 4· Electrostatic and Magnetic Deflection, Electrostatic Focusing

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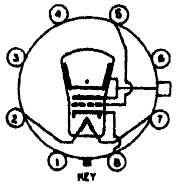
APPENDIX II

RMA CATHODE-RAY-TUBE BASING

4AE



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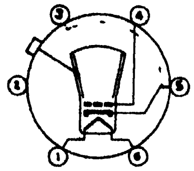

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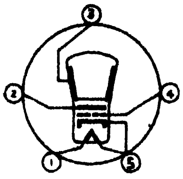

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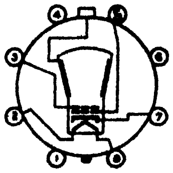

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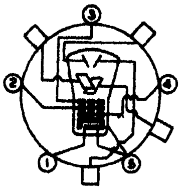

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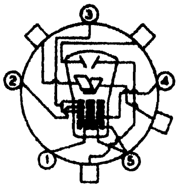

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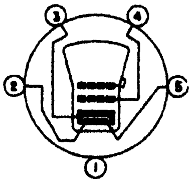

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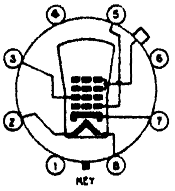

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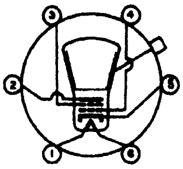

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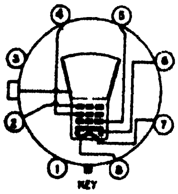

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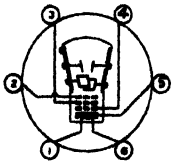

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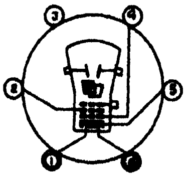

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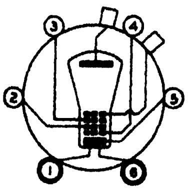

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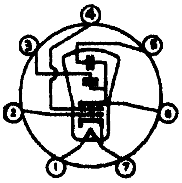

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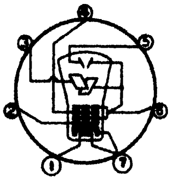

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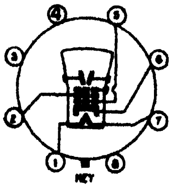

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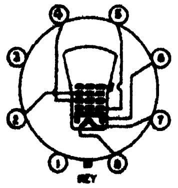

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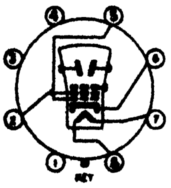

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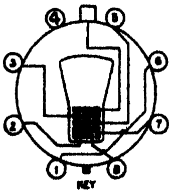

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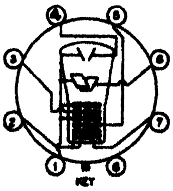

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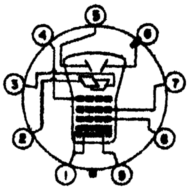

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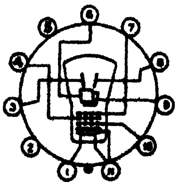

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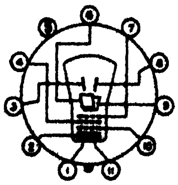

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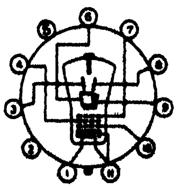

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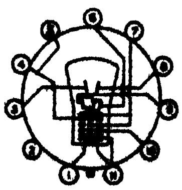

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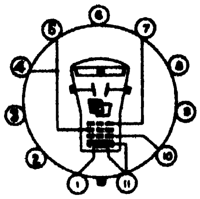

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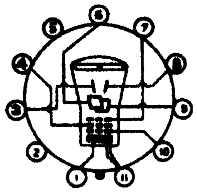

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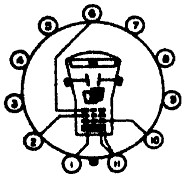

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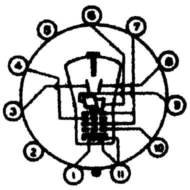

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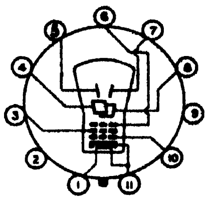

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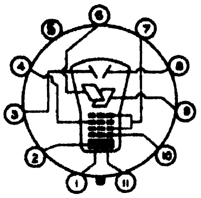

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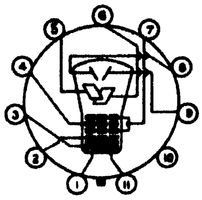

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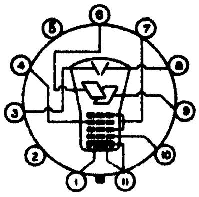

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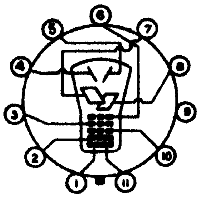

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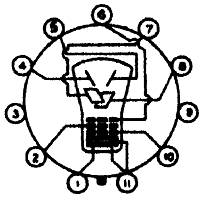

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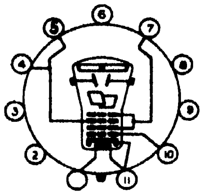

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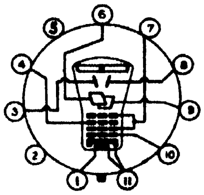

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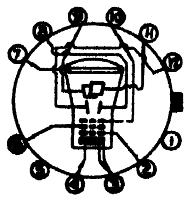

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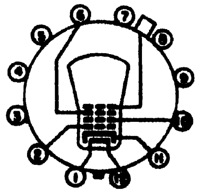

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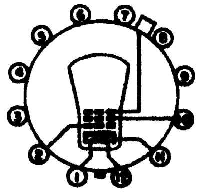

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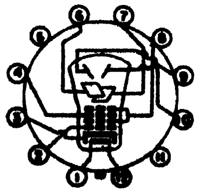

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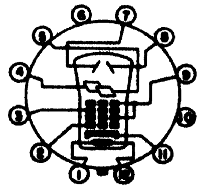

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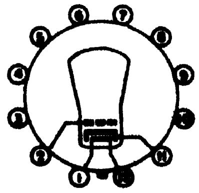

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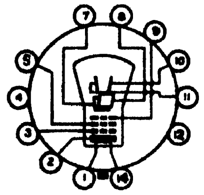

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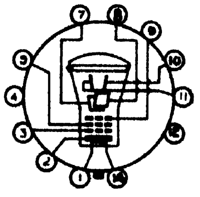

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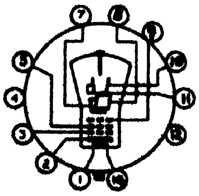

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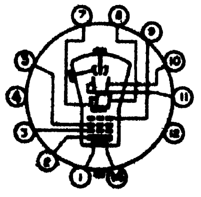

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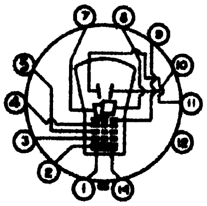

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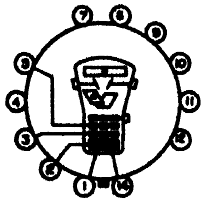

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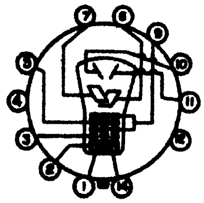

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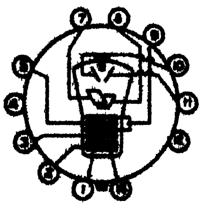

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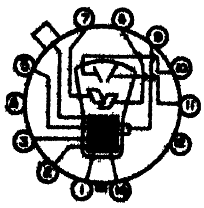

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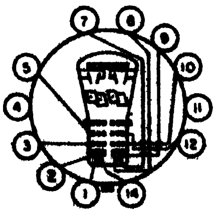

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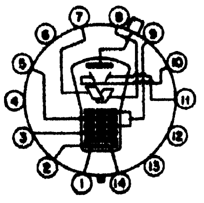


APPENDIX III

PHOTOGRAPHY

This section will begin with a brief discussion of methods of photographing simple oscilloscopic phenomena, to be followed by a more detailed discussion of the more complicated phenomena.

Simple Methods

The simplest, if not the most convenient, method of photographing images on the screens of oscilloscope or television tubes is in a dark room. This may be accomplished if not too much auxiliary equipment is necessary or if long cables may be used.

Usually, however, it is found to be more convenient to employ such methods as are illustrated in Figs. 1 and 2. Here, it will be seen, the equipment used is extremely simple and inexpensive. Moreover, no stringent requirements must be met by the camera, provided it is fast enough to catch whatever nonstationary patterns it may be called upon to record.

An even simpler arrangement, which may be used if lighting conditions are satisfactory, is shown in Fig. 3.

All of these methods have the disadvantage of requiring either an additional shot for recording the necessary data, or some inconvenient way of keeping track of the waveforms, and affixing the date data afterward. A method of photographing both the waveform and the corresponding data simultaneously[1] is given in Fig. 4. This system is quite flexible; by varying the distances d_1 and d_2 (A) and by controlling the intensity of the fluorescent bulb (with a variable autotransformer, for example), a variety of conditions may be accommodated.

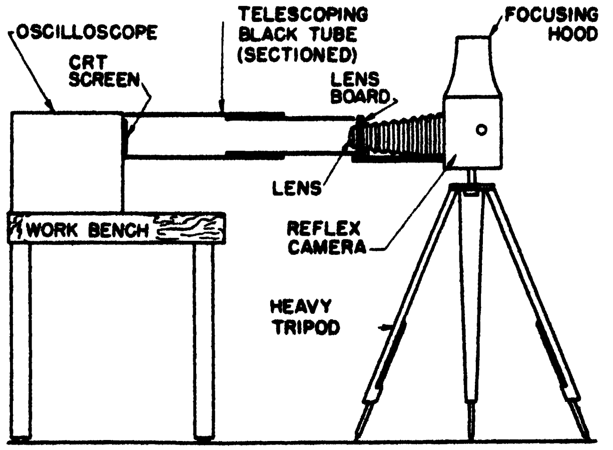


Fig. 1.—Easy method for photographing simple oscilloscopic phenomena.

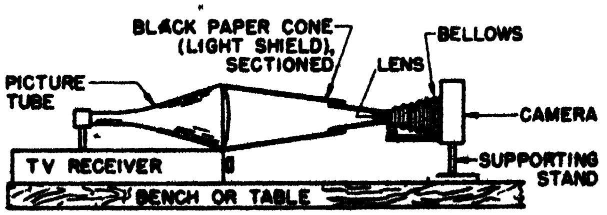


Fig. 2.—Another simple method for oscilloscope photography.

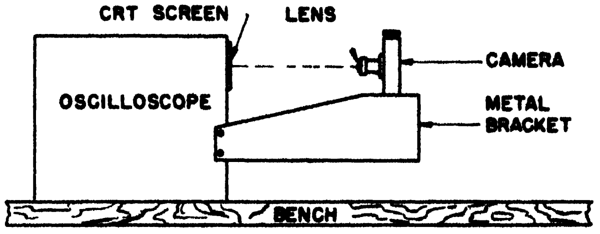
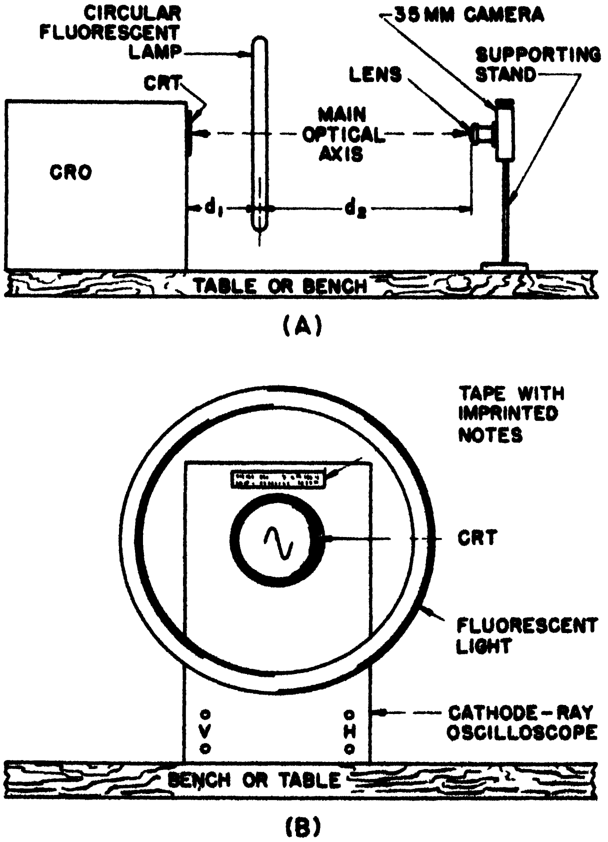


Fig. 3.—One of the simplest arrangements for oscilloscope photography.



Courtesy Electronics

Fig. 4.—Method for simultaneous photographing of both the waveform and the corresponding data.

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[1] Fulmer, Norman, "Improved c-r photographs," *Electronics*, vol. 23, pp. 86-87, March, 1950.

DU MONT PHOTOGRAPHIC DATA[2]

The Du Mont Type 314 Camera was developed for the specific purpose of photographing the patterns on cathode-ray oscilloscopes. In the fields of electricity and electronics, physics, biology, physiology, mechanics, hydraulics, geology, and acoustics, it is often necessary to obtain permanent records which can be studied at convenient times or which must be maintained for reference.

Three general methods may be used to obtain moving-film recordings with the Type 314. The first utilizes the motion of the film as a time base and records along the length of the film. The phenomenon to be recorded is connected to the oscilloscope to produce deflection at right angles to the motion of the film. Since the film moves vertically in front of the cathode-ray tube, this means that the signal must either be connected to the horizontal amplifier of the oscilloscope or to the vertical amplifier, with deflection-plate connections interchanged so that horizontal deflection is produced. On many Du Mont oscilloscopes, deflection-plate connections may be changed by a strap and terminal arrangement at the rear without removing the oscilloscope from its cabinet. On other oscilloscopes, the same result can be achieved by reversing connections internally, or by rotating the cathode-ray tube 90°. It is ordinarily preferable to use the vertical amplifier of the oscilloscope because of its greater sensitivity and wide-band frequency response. However, in some cathode-ray oscilloscopes, where vertical and horizontal amplifiers are identical, there is no preference. This method is entirely satisfactory for relatively slow- or medium-speed phenomena, since the speed which can be recorded is limited by the film speed and resolution of the film. Frequencies up to about 10,000 cps can be handled by this method with the film speeds available in the Type 314.

The second method utilizes both the oscilloscope sweep and film motion to record each successive sweep across the width of the film. With this method the film need travel only at a rate sufficient to separate the sweeps; the only limitation on the speed which can be recorded in this manner is the maximum photographic writing speed of the oscilloscope itself.

The third method is useful in photographing phenomena which begin with frequency components much higher than those which occur thereafter. A single sweep on the oscilloscope connected to move the fluorescent spot in the same direction as the film motion may be used to spread out the initial portion of the phenomenon over a greater length of film. When the sweep is finished, the film motion alone provides the time base.

It is also possible to operate the Type 314 in a manner similar to that of a drum camera by splicing together the two ends of a short length of film and threading the resulting film loop into the camera. The camera can then be set to run at any speed and the loop of film will run continuously for as long as desired.

With this particular method of recording, the camera can serve as a high-speed monitoring recorder and offers a solution to the problem of photographing a transient which occurs at an unpredictable moment.

Photographing stationary patterns produced on the cathode-ray-tube screen by recurrent signals usually is a simple matter.

Courtesy Du Mont Labs.

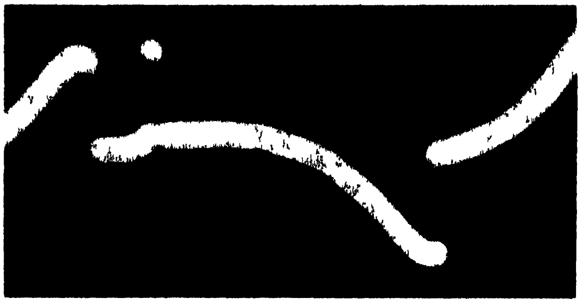
Fig. 5.—The shutter must be left open for at least the time of one complete cycle or else only part of the pattern will appear as indicated here.

Almost any camera which can be focused at close range can be used, but since the image on the film must be unaffected by external light, the exposures have to be made with the oscilloscope and camera in a darkened room. To avoid this inconvenience to the operator each time he has to make a recording, special oscilloscope cameras are available which are provided with light shields and fixed focus so that any cathode-ray-tube pattern can be photographed quickly and conveniently.

To photograph stationary patterns, the camera shutter need be left open only long enough to obtain the required negative density. However, there are several difficulties that may be encountered, of which one should be aware.

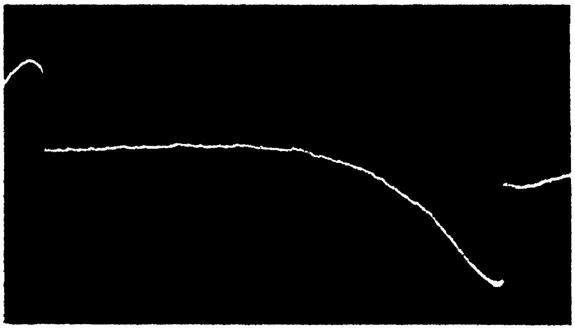
For example, if the shutter is not left open for at least the time of one complete sweep cycle, part of the pattern will be missing, as shown in Fig. 5. It is best to use an exposure time at last several times longer than the time of one sweep cycle to obtain uniform density over the entire portion of the image.

With some types of signals (such as square waves) where the writing rate over various portions of the cycle changes greatly, with resultant large variations in brightness over different parts of the pattern, it may become necessary to overexpose the brighter parts of the pattern in order to obtain satisfactory recording of the less intense portions.



Courtesy Du Mont Labs.

Fig. 6.—Film was deliberately over-exposed to record extremely rapid portions of the trace.



Courtesy Du Mont Labs.

Fig. 7.—A similar film, reduced by chemical means. Note improvement in clarity of detail.

Figs. 6 and 7 illustrate this technique. It is nearly always preferable to overexpose if one is doubtful about the exposure, because the negative always can be reduced by conventional chemical means. In fact, chemical reduction of the oscillogram is often used in the laboratory as a means of enhancing the contrast between the trace and background fog.

Often it is desirable or necessary to use color filters between the cathode-ray-tube screen and camera. A filter having a color transmission the same as the fluorescent trace may improve contrast by darkening the screen background, particularly when used with long-persistence screens.

There are many techniques in oscilloscope photography which enhance the use of oscilloscopic recordings as a means of obtaining data. For example, any number of traces could

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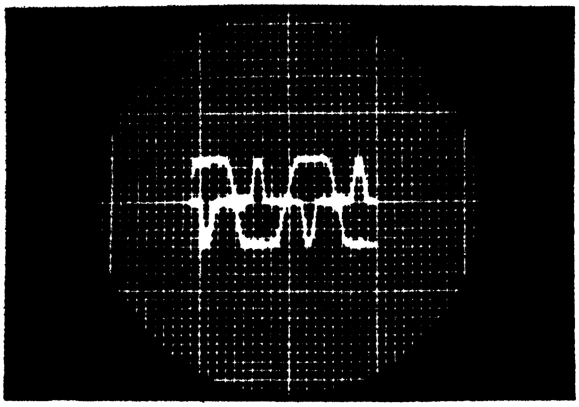
[2] The following material is reprinted (with some portions edited) from the Du Mont "Catalog of Equipment for Oscillography," 1949 edition.



Courtesy Du Mont Labs

Fig. 8.—A triple exposure photograph. (For other triple exposure photographs see Chapter 12.)

be superimposed upon one another for later comparison, merely by making multiple exposures. The oscillogram in Fig. 8 shows a calibration signal from a Du Mont Type 264-B Voltage Calibrator superimposed upon the clipped peaks of a sine wave. Three exposures were made in this case: one of the sine-wave peaks, another of the flat-topped calibrating waveform, and a third of an edge-illuminated engraved lucite scale. Note how an oscillogram such as this immediately makes a true graph recorder out of a conventional cathode-ray oscilloscope. Multiple exposures also make it possible to record data on the same negative as the trace recording. All that is needed is an illuminated ground glass surface upon which one can write data, and a suitable mask to prevent fogging the film. The oscillogram of Fig. 9 was obtained by first making an exposure of the trace and then placing the camera on an illuminated box with a ground glass on which the data were inscribed with an ordinary lead pencil, making a double exposure.

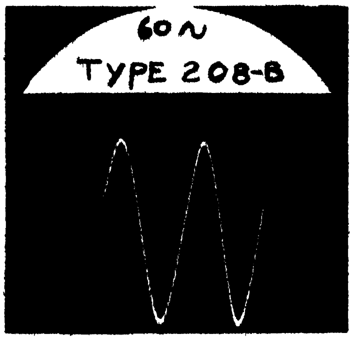


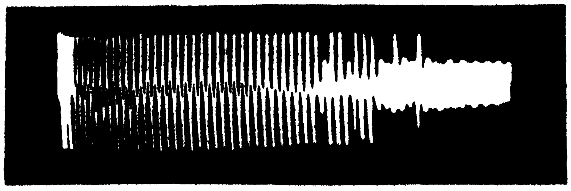
Fig. 9.—Double exposure including waveform and necessary data.

Courtesy Du Mont Labs.

Recording by Means of Continuous-Motion Cameras

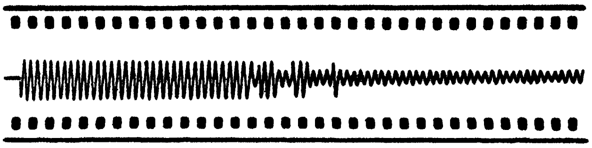
There are two general methods applicable to photographic recording of cathode-ray-tube patterns—a moving-film method, and a stationary-film method. The former method (usually called continuous-motion recording) is preferable when the duration of the observed signal is longer than that of the longest sweep available on the oscilloscope, or when it is desirable to expand a transient in its entirety for more accurate study. For example, Figs. 10 and 11 are recordings of the same phenomena, that of the starting-current characteristic of a synchronous motor. The oscillogram in Fig. 10 was made by a single exposure on stationary film, with a Du Mont Type 271-A Camera, and using a very slow oscilloscope sweep. The oscillogram of Fig. 11 was made on continuously-moving film with a Du Mont Type 314 Camera. It is obvious that while Fig. 10 provides a good over-all record for qualitative study, the expanded oscillogram of Fig. 11 is preferable where an accurate quantitative analysis is necessary.

In the continuous-motion method of recording, several techniques are possible, any of which may be used with the Type 314 Camera.



Courtesy Du Mont Labs

Fig 10.—Single exposure photographic recording of the starting-current characteristic of a synchronous motor.



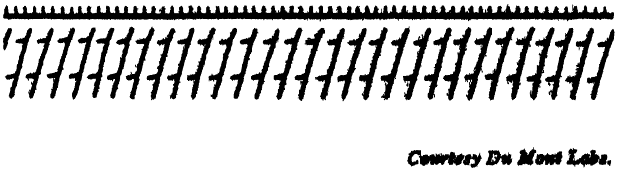
Courtesy Du Mont Labs

Fig 11—Moving film recording of the same starting-current characteristics as shown in Fig 10

In the first method, the spot on the cathode-ray tube is deflected by the signal along one axis only, and the time axis is provided by the motion of the film in the direction perpendicular to the spot deflection This method has certain limitations. In order to obtain sufficient resolution on ordinary film, the recording of a 100-kc signal, for instance, would require a film speed of at least 50 feet per second This method leads to considerable film consumption and is therefore recommended only for the recording of low- and medium-frequency signals.

In cases where it is necessary to provide a monitor for signals occurring completely at random over long periods of time, a continuous-motion recorder is, of course, mandatory. The Du Mont Type 314 Camera, for example, may be used at its lowest speed to record continuously for a period of 8⅓ days.

A second method of continuous-motion recording, which has advantages in many cases, is to use a driven or recurrent sweep on the cathode-ray oscilloscope and to move the film perpendicularly to the direction of the sweep at a speed determined by the repetition rate of the sweep and the amplitude of the recorded signal. This makes it possible to obtain recordings of very high frequencies or of very short pulses with a relatively small amount of film, and it facilitates the analysis of the resulting oscillograms. Both methods may be compared on Fig. 12 which shows at the top a recording by method 1 and at the bottom a recording by method 2. Both recordings are of the same signal. This photograph was obtained by running the same film strip twice through the camera, exposing only part



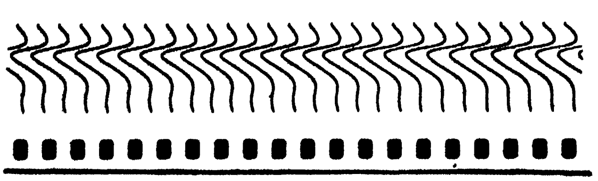
Courtesy Du Mont Labs.

Fig. 12.—A double-exposed strip showing two methods of recording by means of continuous-motion film. The first method (top) records along the length of the film making use of the film motion as the time base. The second method (bottom) utilizes both film motion and oscillograph sweep to record successive sweeps across the width of the film.

of it each time. Obviously, in this case, the second method gives a much clearer picture of what is happening on a considerably shorter length of film.

In recording certain types of phenomena by this method, the pattern may be such that it is possible to allow the successive signals to appear inside one another or "interlace" as is shown in Fig. 13. A much slower recording speed is then permissible.

To record oscillograms of very rapid transients superimposed upon slowly changing signals with sufficient detail for study, it would be necessary, using the first method, to run the film at two speeds, extremely high and extremely low, for the respective components. This procedure is impractical because it is virtually impossible to change the speed of the film at the right instant.



Courtesy Du Mont Labs

Fig. 13.—Continuous-motion recording illustrating the "interlace" which is possible with certain types of signals. A slow film speed is employed in plotting sweeps across the film width

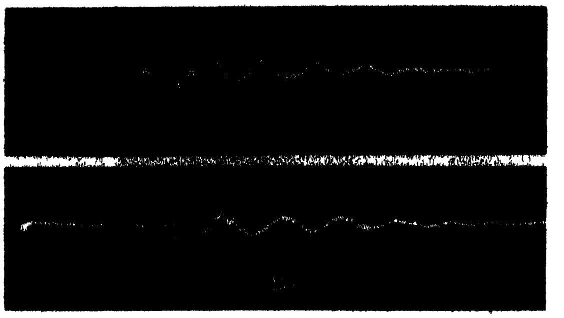
The second method described for recording continuous motion is not always suitable for this purpose, either, since the consecutive time bases would cause too much discontinuity in the recording.

A third technique of recording includes a combination of the oscilloscope sweep speed and the speed of the film to provide a complex time base, which is faster or slower than either the sweep or the film motion (depending on the direction of each motion).

For example, in one application, it was necessary to record a phenomenon which began as a rapid complex transient and concluded as a slowly changing signal. The transient initiated a single-sweep circuit in the cathode-ray oscilloscope while the film was moving slowly; the direction of the sweep as referred to the film plane was opposite to the motion of the film. Thus, the speed of the resulting time base was equal to the sum of the speeds of the film and the optically reduced sweep. For the remainder of the record, it was equal to the speed of the film alone. By providing sweeps triggered by the transients and moving in a direction opposite to the direction of the motion of the film (as referred to the screen of the cathode-ray tube), it is possible to run the film at a low speed and still provide a very fast time base. Since the motion of the film is vertically upward in the 314 Camera, the spot at the beginning of each sweep period would have to be at the top of the screen of the cathode-ray tube and the spot must be deflected downward rapidly by each transient. The signal produces horizontal deflection of the spot which must remain unblanked at the end of its motion. With the use of an exponential sweep rather than a linear sweep, a discontinuity in the time base can be avoided; that is, the exponential curve of the sweep rate can be made asymptotic to the curve of the speed of the film. With this method of recording, the use of timing markers is necessary in order to achieve the proper time perspective when studying the recording.

A system of recording a time-marker track along the edge of the film, as is provided by the Type 314 Camera, is suitable. Summing up, then, the time base for the rapid transient is provided by the sum of the optically-reduced sweep speed (transferred to the film plane) and the speed of the film. The time base for the slow rate-of-change portion is provided by the film motion alone.

An example illustrating the usefulness of this recording method is shown in Fig. 14A. This oscillogram was obtained by shocking a loaded coil spring into oscillation. The spring was loaded by suspending a Du Mont Type DP-1 Displacement Pickup from one end and connecting the pickup leads to the oscilloscope. Note the high-frequency shock oscillations at the start of the recording and the low-natural-frequency vibrations of the spring after the shock. Note also how the great difference in writing rates causes the high frequencies to be underexposed while low frequencies are overexposed. Fig. 14B shows the recording that is obtained using the first method. Note that all detail is lost in the high-frequency oscillation, while the low-frequency vibrations are displayed satisfactorily.



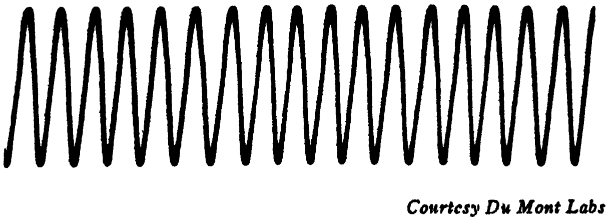
Courtesy Du Mont Labs.

Fig. 14.—(A) Shock transient and subsequent oscillations of coiled spring, recorded using both oscillograph sweep and film motion; (B) same as Part (A) without the use of the oscillograph sweep. Note how initial detail of trace is lost.

For all recordings by means of continuous-motion cameras, regardless of the method used, only short-persistence screens can be employed to avoid blurring of the film by afterglow. The seriousness of this effect is shown by Fig. 15, which represents the recording of a 60-cps signal on a medium-persistent P1 screen. For comparison, the recording of a 1,000-cps signal on a short-persistence P11 screen may be seen on Fig. 16. No blurring can be observed on this recording.

Courtesy Du Mont Labs.

Fig. 15.—Continuous-motion recording of a 60-cps sine wave. Blurring is due to the long-persistence characteristic of screen material.

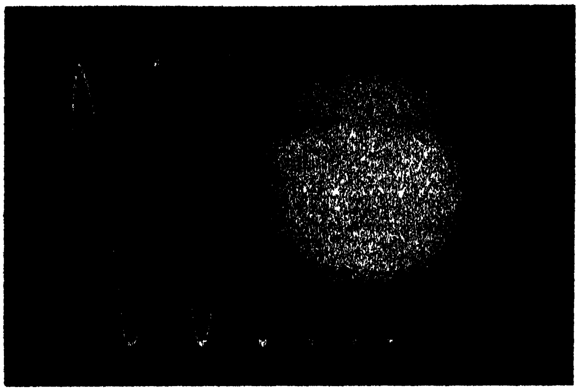


Courtesy Du Mont Labs.

Fig. 16.—Continuous-motion recording of 1000-cps sine wave from a short-persistence, P11 screen. Note clarity as compared to Fig. 15.

Extremely high recording speeds may be obtained with limited film consumption by means of the so-called "drum camera." This camera contains a cylindrical drum on the circumference of which a strip of film or sensitive paper is fastened. The drum is driven to full speed, then the shutter opens for one revolution, while the signal is simultaneously applied to the cathode-ray tube.

The Type 314 Camera may be operated in a manner somewhat similar to that of a "drum camera" by loading a continuous loop of film (obtained by splicing together the ends of a length of film) into the camera. This technique has advantages in certain applications where it is desirable to run the camera at maximum speed for an indefinite time in order to record a random transient of very short duration. The film must be kept from fogging by using a cathode-ray oscilloscope having automatic beam control.



Courtesy Du Mont Labs.

Fig. 17.—Fogging due to cathode glow obscures much of the trace.

Recording of Single and Repetitive Transients

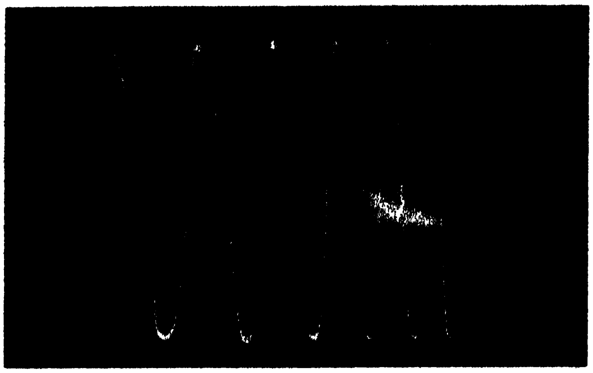
The photographic recording of single transients from cathode-ray-tube screens requires considerably more care and technique if satisfactory results are to be obtained. To photograph single transients, two basic methods may be used:

1. The shutter may be opened automatically shortly before, or exactly when the transient occurs and closed immediately after.

2. The shutter may be left open on "Bulb" or "Time" until the transient occurs after which the shutter is closed. In photographing single transients, two basic methods may be used.

The latter method is generally used because it is accomplished more easily. In recording transients, it is particularly important to avoid background light, due to extraneous signals and stray electron excitation of the screen which might mask the signal or reduce the contrast of a low-density recording. Overexposure by a spot or line should be avoided by the use of oscilloscopes having driven sweeps and automatic beam-blanking circuits so that the beam is on only when the transient occurs.

One difficulty that is sometimes experienced is that of a cathode-glow (light coming from the cathode-ray-tube heater) penetrating the screen and fogging the film. This occurs with a sensitive panchromatic emulsion, when the shutter is left open for a considerable time before or after the transient occurs. An example of this difficulty is shown in the oscillogram of a 20-Mc single transient, Fig. 17. Note that the fog masks almost the entire signal. Fortunately, this fogging can be reduced or eliminated by several methods:



Courtesy Du Mont Labs

Fig. 18.—Pattern similar to Fig. 17. Cathode glow is reduced by shortening the time that the shutter is opened before and after the transient occurs.

(a) A nonred sensitive emulsion such as Eastman Linagraph Ortho (5211) or Ansco Triple S Ortho may be used with some reduction in emulsion speed as compared to Eastman Linagraph Pan (5244).

(b) A blue filter, such as the Du Mont Type 216-H or Corning Glass #5030, may be placed before a P11 tube to absorb the red-orange light from the filament, with some loss in transmission of the desired light.

(c) The first-mentioned method of photographing transients may be used, in which the shutter is either manually or automatically opened shortly before the transient occurs and closed as quickly as possible after the transient is over. This method avoids loss of exposure, although the fogging is not always completely eliminated, as shown in the oscillogram of Fig. 18. This oscillogram is of the same pattern as Fig. 17.

The density obtained determines the maximum speed transient which can be obtained with any given oscilloscope, camera, and film emulsion.

In photographing repetitive transients, the same considerations are used as are used for recurrent signals and stationary patterns. For a given sweep speed, the exposure obtained will be proportional to the repetition rate of the transient.

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